RADIO FREQUENCY CIRCUIT DESIGN FOR AUTOMATIC DATA PROCESSING OF RADAR SIGNALS OF REMOTE SEA-ICE THICKNESS SENSOR

Final Report

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NOLOGIES INC.
Road, Trans Canada Highway
ue, Quebec, Canada H9X 3L5

MPB TECHNOLOGIES INC.
P.O. Box 160
21051 North Service Road | Trans-Canada Highway
Ste-Anne-de-Bellevue, Québec H9X 3L5 | Téléphone (514) 457-2035

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RADIO FREQUENCY CIRCUIT DESIGN FOR AUTOMATIC DATA PROCESSING OF RADAR SIGNALS OF REMOTE SEA-ICE THICKNESS SENSOR

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Final Report

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An English summary of this report is included immediately preceeding the Table of Contents.

Prepared for:

Transport Canada
Transportation Development Centre (TDC)
1000 Sherbrooke Street West
P.O. Box 549
Montreal, Quebec
H3A 2R3

Prepared by:

MPB Technologies Inc.
21051 North Service Road
P.O. Box 160
Trans-Canada Highway
Ste. Anne de Bellevue, Quebec
H9X 3L5

G. Grandchamp and S. Geller

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Approved: M P Bachynski

Dr. M.P. Bachynski
President

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17. Abstract

This report presents theoretical and practical considerations on the feasibility of elements believed to be the most critical of the sea-ice thickness sensor radio frequency (R-F) section. These elements are the sample and hold units, the transmitter R-F power amplifier and the receiver R-F switch. This report also proposes structural changes in the R-F section, such as elimination of the even-numbered frequencies from the synthetic pulse and elimination of the variable capacitance diode phase shifters, which will lead to hardware simplification. The results of this study will be directly applied to the R-F designs of the sensor developmental prototype.

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17. Résumé

Ce rapport présente les résultats d'une analyse théorique et pratique des possibilités de réalisation des circuits les plus critiques de la haute fréquence d'un système de télédétection de l'épaisseur de la glace de mer. Ces éléments comprennent le dispositif d'échantillonnage et maintien, l'amplificateur de puissance haute fréquence (HF) de l'émetteur et le commutateur HF du récepteur. Le rapport suggère aussi de simplifier la structure de l'émetteur et du récepteur en ne conservant que les composantes harmoniques d'ordre impair de l'impulsion de synthèse et en évitant l'utilisation de diodes à capacité variable pour la réalisation des déphaseurs. Les résultats de cette étude seront mis en pratique lors de la réalisation de la maquette de développement du système de télédétection.

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Glace de mer, système de télédétection, radar, haute fréquence, très haute fréquence, échantillonnage, commutation	Disponibilité d'exemplaires en nombre limité au CDT

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PROJECT TEAM

The project team for MPB Technologies Inc. was composed of:

Mr. G. Grandchamp
Mr. S. Geller

SUMMARY

1. SCOPE

This document summarizes the work carried out at MPB Technologies Inc. on Contract 0SD80-00128 between March 15, 1981 and October 25, 1981.

The objective was to investigate the feasibility of the most critical elements of the sea-ice thickness sensor RF section which will be needed for the development of a fully automated sensor. Prior to the start of this contract, these elements have been identified as being respectively:

- The sample and hold unit used for digitizing the received signal.
- The RF power amplifier needed for boosting up the transmitter power.
- The RF switches used in the transmit-receive unit to prevent the transmitted power from leaking directly into the receiver.

The basic requirements leading to the selection of specific configurations for the sample and hold unit, the RF power amplifiers and the RF switches are briefly reviewed below.

2. SAMPLE AND HOLD UNIT

A sample and hold unit is normally used to convert an analog signal into a sequence of constant levels of identical duration. These levels are then converted into a bit sequence by an analog-to-digital converter, so that the signal can be stored and processed digitally. A sampling unit consists of an analog switch, periodically closed by a gating signal. The input signal is then transmitted through the switch during a short time called the sampling interval. The hold unit consists of an amplifier capable of memorizing for some time the value taken by the input signal at the end of the sampling interval.

2.1 Sampling modes

There are basically two modes of sampling

- sampling of aperiodic signals
- sampling of periodic signals

Aperiodic signals are characterized by a continuous frequency spectrum. Sampling of this type of signal requires that the sampling frequency fs should be related to the signal bandwidth B by the inequality $fs \geq 2B$. Whenever this inequality is

satisfied, the original signal can be recovered through low pass filtering or digital filtering. Digital processing of the original signal requires that each sample must be converted into a binary number before the next sample is taken. Consequently, the sampling time must be made as short as possible and smaller than the sampling period.

When the signal is periodic with period T, it is not necessary to collect all the required samples during one period only. Instead, one sample can be collected at each period by incrementing the sampling time by a small amount ΔT_s, which is related to the signal bandwidth B and the number of collected samples N by the inequality

$$\Delta T_s \leq \frac{N}{N+1} \cdot \frac{1}{2B}$$

Sampling of periodic signals enables reduction of the speed requirements of the analog-to-digital converter and is a useful method of sampling high frequency periodic signals.

2.2 Application to the sea-ice thickness sensor

The constraints to be placed on a sampling and hold unit for the sea-ice thickness sensor were established and four sampling methods were examined in detail to handle a train of synthetic pulses composed of 5 frequencies (30 MHz, 60 MHz, 90 MHz, 120 MHz and 150 MHz). These methods are:

- direct aperiodic sampling
- stretched pulse aperiodic sampling
- direct period sampling
- stretched pulse periodic sampling

The applicability of each of these methods to the sea-ice thickness sensor is now briefly reviewed.

. Direct aperiodic sampling of the synthetic pulse train is the most attractive method, but it cannot be implemented in practice because the time allowed for digital conversion is only about 3 ms and no existing analog-to-digital converter is capable of such a high speed.

. One way to avoid the difficulties associated with direct sampling and still retain the concept of aperiodic sampling is to implement a time stretching circuit to convert the synthetic pulse train into a slower synthetic pulse train of smaller bandwidth. The report shows that a minimum bandwidth of 6.6 MHz is achievable. Aperiodic sampling of the stretched pulse train could reduce considerably the speed and bandwidth constraints of the sample and hold unit as well as those of the A/D converter and buffer memories. A sample and hold unit capable of operating conservatively at a sampling frequency of 15 MHz and having a bandwidth of 50 MHz has been procured from "DATA DEVICES CORPORATION" and evaluated. The results show

that this unit would introduce very small attenuation and phase dispersion over the 6.6 MHz bandwidth of the stretched pulse train. Aperiodic sampling offers the advantage of allowing the acquisition of all required data during one repetition period of the sensor transmitted signal. Therefore, the motion of the helicopter or the aircraft carrying the sensor has practically no effect on the received data. The disadvantage is that a large number of synthetic pulses is required and this puts a constraint on the altitude from which the sensor must operate to avoid interferences between the transmitted and received signals. Under the minimum 6.6 MHz bandwidth conditions, 100 synthetic pulses are required in the original pulse trains, leading to a pulse train duration of 3.3 ms. The helicopter or the aircraft would then have to operate at an altitude of 470 metres and the power requirements would have to be increased accordingly.

. Direct periodic sampling assures that the return signal from the ice is stationary during the whole acquisition interval. This is true if the sensor repetition frequency can be made large enough so that the helicopter or the aircraft only moves by a negligible amount during the acquisition interval. For the present frequency plan and for a train of 6 synthetic pulses, transmitted with a repetition frequency of 50 KHz, the acquisition time for 6400 samples is shown to take place in 128 milliseconds for a sampling time increment of 3 ns. During that time, a helicopter moving horizontally at 50 km/h would only change its position by 2.5 metres and the interpretation error is likely to be quite small. Therefore, direct periodic sampling is very attractive and it is certainly the solution to adopt if the high frequency problems associated with the short sampling interval and the scanner time jitter can be solved.

. Stretched pulse sampling can also be used for periodic signals. However, in this case, the advantages of the pulse stretching techniques described for the aperiodic case are partly lost, because there is little need to lower the sampling frequency. However, the method would still be useful to reduce the bandwidth requirements of the sample and hold unit if the direct periodic sampling method were not satisfactory. It must be noted that the resulting scheme would increase the minimum altitude at which the helicopter or the aircraft would have to operate. This increase would depend on the desired bandwidth reduction.

2.3 Future Objectives

A comparison of the advantages and disadvantages of the four sampling techniques shows that the direct periodic sampling method is the best suited to the sea-ice thickness sensor for the following reasons:

- No limitations of the minimum operating altitude of the sensor
- No frequency conversion required
- Only low speed A/D converters are necessary
- Critical high frequency operation limited to sampling gate and associated drive circuitry

Consequently, it has been decided that a direct periodic sample and hold unit and associated scanning circuitry will be developed during the RF section design phase of the sensor developmental prototype.

3. POWER AMPLIFIER

The peak power required at the sea-ice thickness sensor transmitter output is a function of the peak signal-to-noise ratio tolerable at the receiver input. It is commonly understood that a peak signal-to-noise ratio of 20 to 30 dB at the receiver will allow adequate resolution of the received signal. For a given noise level at the receiver input, the peak signal power required at the transmitter output will then be a function of the following parameters:

a) operating altitude
b) reflection coefficients for the upper and lower sea-ice boundaries
c) Antenna gain characteristics

The helicopter operating altitude will likely not exceed 300 metres. The reflection coefficients for the ice upper and lower boundaries have known limits for the sea-ice thicknesses under consideration in this project.

The gain characteristics of the newly developed antenna are under current evaluation and will not be known accurately before some form of trial in the field takes place.

The antenna unknown characteristics, coupled to the fact that the noise level at the receiver input can vary within very large limits depending on where the sensor is operated, make it very difficult to predict any transmitter power requirements at the moment. In the arctic sea, the noise is predominantly thermal and generated in the antenna and receiver electronics. In other regions which might be selected to carry out preliminary trials, the electromagnetic interferences from nearby VHF transmitters become the predominant source of noise. Under these conditions, it was decided to increase the maximum peak power of the transmitter to such a level that even the worst interferences would not have any significant effects on the received signal interpretation. Based on previous measurements made on the sea-ice thickness sensor experimental prototype, it was felt that a transmitter peak power ranging from 2 to 10 watts would be adequate. Design considerations on solid state power amplifiers and some of the problems encountered when trying to amplify synthetic pulses are summarized here.

3.1 Power Amplifier Design Considerations

The amplification of the synthetic pulse train at the transmitter end can be achieved in two ways:

- Direct amplification of the synthetic pulse
- Separate amplification of individual frequency components of the synthetic pulse.

These two methods will now be briefly reviewed.

. In the Direct Amplification Mode, the synthetic pulse is amplified as linearly as possible through a broadband VHF power transistor amplifier operating in a class A mode. The connections to the antenna must be made through some form of isolating device to prevent excessive reflected power from the load from reaching the output transistor stage. Two cases are possible depending on where switching takes place.

  - If switching is done between the amplifier output and the antenna, the transmitter switch could be embedded in an absorptive type of arrangement to minimize reflections from the antenna.

  - If switching is done before the power amplifier and if the antenna mismatch is kept small (VSWR < 2), the antenna can be connected directly, but the amplifier must be capable of transmitting the switched pulse train without significant rise time and fall time degradations.

In both cases, the power amplifier must have the following characteristics:

a) low harmonic generation
b) low amplitude and phase dispersion with frequency
c) clean transient response
d) reasonable efficiency

A linear power amplifier presumably supposed to meet these requirements has been procured from "MICROWAVE POWER DEVICES INC." (type EWA-2015) and tested in conjunction with the sea-ice thickness sensor experimental prototype. The amplifier is capable of operating over the frequency range 20 MHz to 150 MHz with a ± 0.5 dB gain variation. The amplifier nominal CW output power is 10W for an input power of 1 mW. The transient response is satisfactory with a 12 ns rise time and a 25 ns fall time. The efficiency is low and of the order of 10%. The harmonic generation was found to be -22 dB for the third harmonic. This level of third harmonic generation is not satisfactory, because it will impair the adjustment of the synthetic pulse shape. The synthetic pulse is made up of harmonically related frequency components with independent amplitude and phases. Excessive intermodulation within the power amplifier will make independent amplitude and phase control impossible. A detailed study of the influence of harmonic generation on the synthetic pulse shape has not been undertaken yet, but it is likely that a level of -30 dB to -40 dB on all harmonics will be necessary. Therefore, the power amplifier presently available does not appear fully suitable for synthetic pulse amplification.

. In the Separate Amplification Mode, the intermodulation problems encountered in the previous method can be eliminated by amplifying each frequency component in separate narrowband class C amplifiers. The individual signals are then linearly added at the output. Transistor class C power amplifiers have also a better efficiency than class A amplifiers but the switching speed is affected by the narrower bandwidth. Under these conditions, switching will have to be performed between the power amplifier and the antenna.

3.2 Future Objectives

No clear decision has been made yet on the amplification mode which will be adopted for the developmental prototype. Although the measurements made on an existing power amplifier unit in the direct amplification mode were not fully satisfactory from the intermodulation standpoint, this mode of amplification still has the advantage of allowing switching to take place at low levels, before the amplifier. The power leak to the receiver is thus likely to be smaller than in the separate amplification mode and the pulse rise and fall times will be better preserved. Consequently, it has been decided to take a second look at the direct amplification mode during the design phase of the RF sections.

4. RF SWITCH

A basic transmit-receive (TR) switch configuration includes:

- A transmitter switch, used to gate the synthetic pulse train and prevent, if necessary, excessive reflected power from the switching devices and the antenna from interacting with the power amplifier.

- A receiver switch used to prevent the transmitter from leaking into the receiver during transmit time.

Both switches must meet the following constraints:

a) sufficient isolation
b) low insertion loss
c) suitable power handling capability
d) low intermodulation, compatible with synthetic pulse shape preservation
e) low switching transients
f) switching speed compatible with the minimum operating altitude of the sensor.

Except for the power handling capability, the receiver switch is more critical than the transmitter switch. Consequently, only the receiver switch has been investigated within the frame of the present contract. Measurements made on the receiver switch used in the sea-ice thickness sensor experimental prototype are given in the report and alternative switch configurations suitable for the developmental prototype are discussed.

4.1 Switch Evaluation

A complete evaluation of this switch was made within the frame of this contract. The following comments were made:

- The switch isolation for a 4 mW input signal varying in frequency from 30 MHz to 150 MHz is at least -60 dB. This is still above the level of the expected attenuation of the sensor signal path and is just acceptable.

- The insertion loss is about 4 dB for a 10 mA driving current. This is acceptable.

- The power handling capability of the switch in the off state is limited to approximately 15 mW. This is due to the fact that no back-biasing of the diodes is possible when the switch is in the off state and therefore the switch can be easily driven into the on state when the RF signal exceeds a certain power level.

- The second and third harmonic generation in the on state are below -30 dB for a 4 mW input power level. This is compatible with good preservation of the synthetic pulse shape.

- The switching transients, although already reduced by the circuit symmetry, are in excess of 20 mV for a 100 mV driving pulse across the switches. The spikes duration is of the order of 7 nanoseconds. This level of transients has been found to overload the receiver and create unacceptable ringing during receiving time.

- The switching speed is of the order of 20 ns.

4.2 Future Objectives

The previous results lead to the conclusions that the switch does not have the power handling capability required for the developmental prototype and that the switching transients are still too strong. Although the transient level has been reduced by the symmetric arrangement described in the report, the power handling capability still needs improvements. No modifications can be made to the present switch, because the transformers and the diodes are not accessible. Therefore, it has been decided to review the design of the T-R switch for the developmental prototype. Although no realization has been made yet, design considerations for a new switch are presented in the report.

RÉSUMÉ

1. APERÇU

Ce document résume le travail effectué aux "Technologies M.P.B. Inc." sur le contrat OSD80 - 00128 du 15 mars 1981 au 25 octobre 1981.

Le but du contrat était d'étudier les possibilités de réalisation des éléments de circuit les plus critiques de la partie haute fréquence d'un système de télédétection d'épaisseur de la glace en mer. Avant le début du contrat, ces éléments, nécessaires au développement d'un système de télédétection complétement automatique, avaient été identifiés comme étant respectivement

- L'unité d'échantillonnage et maintien servant à la numérisation des signaux reçus.
- L'amplificateur final de puissance de la partie émettrice.
- Les commutateurs haute fréquence de l'unité d'aiguillage émission-réception, utilisés pour éviter que la puissance émise pénètre dans le récepteur.

Les exigences fondamentales conduisant à la sélection d'une configuration précise pour l'unité d'échantillonnage et de maintien, l'amplificateur de puissance et les commutateurs haute fréquence sont brièvement résumées ci-dessous.

2. UNITE D'ECHANTILLONNAGE ET DE MAINTIEN

Une unité d'échantillonnage et de maintien est normalement utilisée pour transformer un signal analogique en une suite de niveaux constants d'égales durées. Ces niveaux sont ensuite convertis en une suite binaire à l'aide d'un convertisseur analogique - numérique (A/N), de manière à pouvoir conserver le signal en mémoire et le traiter ultérieurement. Une unité d'échantillonnage se compose d'un commutateur analogique excité périodiquement à l'aide d'un signal de pilotage; le signal d'entrée est alors transmis au travers du commutateur pendant un temps très court, appelé intervalle d'échantillonnage. L'unité de maintien consiste en un amplificateur capable de garder en mémoire la valeur prise par le signal d'entrée à la fin de l'intervalle d'échantillonnage.

2.1 Mode d'échantillonnage

Il existe deux modes d'échantillonnage

- l'échantillonnage de signaux apériodiques
- l'échantillonnage de signaux périodiques

Un signal apériodique est caractérisé par un spectre continu de fréquence. L'échantillonnage de ce type de signal exige que la fréquence d'échantillonnage fs soit liée à la bande passante B du signal par l'inégalité $fs \geq 2B$. Lorsque cette inégalité est satisfaite, le signal d'origine peut être alors récupéré en faisant passer le signal échantillonné au travers d'un filtre passe-bas ou numérique. Le traitement numérique du signal d'origine exige que chaque échantillon soit converti sous la forme d'un nombre binaire avant que le prochain échantillon soit prélevé. Par conséquent, il est nécessaire choisir le temps d'échantillonnage aussi court que possible et beaucoup plus court que la période d'échantillonnage. Lorsque le signal est périodique de période T, il n'est plus nécessaire de prélever tous les échantillons au cours d'une seule période. Dans ce cas, il suffit de prélever un échantillon à chaque période en donnant un léger accroissement ΔT_s à l'instant d'échantillonnage; cet accroissement est lié à la bande passante du signal et au nombre d'échantillons prélevés N par l'intermédiaire de l'inégalité

$$\Delta T_s \leq \frac{N}{N+1} \cdot \frac{1}{2B} .$$

L'échantillonnage d'un signal périodique permet de diminuer considérablement les exigences de rapidité du convertisseur analogique - numérique; cette méthode est utile lorsqu'il s'agit d'effectuer un échantillonnage de signaux périodiques en haute fréquence.

2.2 Application aux systèmes de télédétection d'épaisseur de la glace en mer

Les spécifications d'une unité d'échantillonnage et de maintien destinée à un système de télédétection d'épaisseur de la glace en mer ont été établies. Quatre méthodes d'échantillonnage d'un train d'impulsions synthétiques, composé de cinq fréquences (30 MHz, 60 MHz, 90 MHz, 120 MHz et 150 MHz), furent ensuite examinées.

Ces méthodes sont les suivantes:

- échantillonnage apériodique direct
- échantillonnage apériodique de l'impulsion étirée
- échantillonnage périodique direct
- échantillonnage périodique de l'impulsion étirée

Nous allons maintenant passer en revue brièvement les possibilités d'application de chacune de ces méthodes au système de télédétection qui nous intéresse.

- L'échantillonnage apériodique directe d'un train d'impulsions synthétiques est certainement la méthode la plus attrayante, mais il n'est pas possible de l'utiliser dans le cas particulier, car l'intervalle de temps reservé à la conversion numérique n'est que d'environ 3ns et aucun convertisseur n'est capable d'une telle vitesse.

- Une méthode, permettant d'éviter les difficultés précédentes tout en conservant les principes de l'échantillonnage apériodique, consiste à construire un circuit d'expansion permettant de transformer le train d'impulsions synthétiques en un train plus lent, à bande passante plus faible. On montre dans le rapport qu'une bande passante minimale de 6,6 MHz peut être obtenue. L'échantillonnage apériodique du train étiré d'impulsions devrait permettre de réduire considérablement les contraintes de vitesse et de bande passante de l'unité d'échantillonnage et de maintien, de même que celle du convertisseur A/N et des mémoires tampon. Une unité d'échantillonnage et de maintien, capable de fonctionner d'une manière fiable à la fréquence d'échantillonnage de 15 MHz et d'accepter des signaux sur une bande passante de 50 MHz, a été commandée chez "DATA DEVICES CORPORATION". L'évaluation de cette unité a montré que l'atténuation et la distortion de phase sur une bande de 6,6 MHz étaient négligeables. L'échantillonnage apériodique offre l'avantage de permettre l'acquisition de toutes les données nécessaires au cours d'une seule période de répétition du signal émis. Dans ces conditions, le mouvement de l'hélicoptère ou de l'avion, à bord duquel se

trouve le système de télédétection, n'aura pas d'effet sur les données reçues. Le désavantage réside dans le fait qu'un très grand nombre d'impulsions synthétiques sont nécessaires, ce qui introduit une limite sur l'altitude à laquelle le système doit fonctionner pour éviter les interférences entre les signaux émis et reçus. Dans le cas d'une bande passante minimale de 6,6 MHz, il faudrait que le train d'impulsions d'origine contienne 100 impulsions synthétiques, ce qui conduirait à une durée totale du train d'impulsions de 3,3 ms. Il serait donc nécessaire que l'hélicoptère ou l'avion opèrent à une altitude 470 mètres, ce qui augmenterait considérablement les exigences sur la puissance émise.

- L'échantillonnage périodique direct suppose que les signaux de retour en provenance de la glace sont stationnaires pendant toute la période d'acquisition des données. Ceci est vrai si la fréquence de repétition du système de télédétection est choisi à une valeur suffisamment élevée, de façon à ce que l'hélicoptère ou l'avion ne se déplace que sur une distance négligeable au cours de l'intervalle d'acquisition. Dans le cas du plan de fréquence actuel, c'est à dire pour un train constitué de 6 impulsions synthéthiques transmises à la fréquence de répétition de 50 KHz, le temps d'acquisition de 6 400 échantillons aura lieu en 128 millisecondes pour un accroissement de l'instant d'échantillonnage de 3 ns. Pendant ce temps, un hélicoptère qui se déplacerait horizontalement à la vitesse de 50 km/h ne verrait sa position changer que de 2.5 mètres; par conséquent l'erreur d'interprétation qui résulte de ce déplacement est très faible. Il s'ensuit que l'échantillonnage périodique direct est une méthode très intéressante et qu'elle constitue certainement la solution à adopter dans le mesure où les problèmes de haute fréquence associés à l'intervalle d'échantillonnage très court et à la gigue du système de balayage peuvent être résolus.

- La méthode d'expansion des impulsions peut également être utilisée en liaison avec des signaux périodiques. Dans ce cas, les avantages de cette méthode mentionnés dans le cas d'un signal apériodique sont en partie perdus, car il n'est plus ici réellement nécessaire de réduire la fréquence d'échantillonnage. Une telle méthode pourrait toutefois s'avérer utile pour réduire la bande passante de l'unité d'échantillonnage et de maintien, au cas où la méthode d'échantillonnage périodique direct n'aurait pas été satisfaisante. Il faut noter en passant que l'utilisation de cette méthode résulterait aussi dans un accroissement de l'altitude minimale à laquelle l'hélicoptère ou l'avion devrait voler. L'accroissement dépendra bien sûr de la réduction de bande passante envisagée.

2.3 Objectifs futurs

La comparaison des avantages et des désavantages des quatre méthodes d'échantillonnage montre que l'échantillonnage périodique direct est la mieux adapté à la réalisation du système envisagé, de télédétection, ceci pour les raisons suivantes:

- aucune limitation dans l'altitude minimale à laquelle le système de télédétection doit fonctionner
- aucune conversion nécessaire de fréquence
- seuls des convertisseurs A/N de faible vitesse sont nécessaires
- le fonctionnement critique en haute fréquence est localisé au niveau de la porte d'échantillonnage et des circuits d'excitation de cette porte.

En conclusion, nous avons décidé de dévolopper une unité d'échantillonnage et de maintien du type périodique direct au cours de la phase de conception de la partie haute fréquence du prototype de développement.

3. AMPLIFICATEUR DE PUISSANCE

La puissance crête nécessaire à la sortie de l'émetteur du système de télédétection dépend du rapport signal à bruit toléré à l'entrée du récepteur. On admet en général qu'un rapport signal à bruit de 20 à 30 dB à l'entrée du récepteur est suffisant pour permettre une résolution correcte des signaux reçus. Pour un niveau de bruit donné à l'entrée du récepteur, la puissance crête nécessaire à la sortie de l'émetteur est fonction des paramètres suivants:

a) altitude de fonctionnement
b) coefficients de réflexion des surfaces supérieures et inférieures de la glace
c) caractéristiques de gain de l'antenne

L'altitude de fonctionnement de l'hélicoptère ne devrait pas dépasser 300 mètres. Les coefficients de réflexion des surfaces supérieures et inférieures de la glace varient dans des limites connues.

Les caractéristiques de gain de l'antenne en cours de développement ne seront pas connues avec précision avant d'avoir effectué des essais dans des conditions réelles de fonctionnement.

Les caractéristiques mal connues de l'antenne et le fait que le niveau de bruit à l'entrée du récepteur peut varier dans de très grandes limites, dépendant de l'endroit où le système fonctionne, résultent en ce qu'il est très difficile de prévoir avec exactitude la puissance d'émission nécessaire. Dans l'océan Arctique, le bruit est à prédominance thermique et est généré essentiellement dans l'antenne et les équipements électroniques du récepteur. Dans d'autres régions, que l'on pourrait être amené à sélectionner pour effectuer des essais préliminaires, les interférences électro-magnétiques en provenance d'émetteur VHF voisins deviennent la source prédominante de bruit. Dans de telles conditions, nous avons pensé qu'il serait prudent d'augmenter le puissance crête de l'émetteur jusqu'à un niveau où la plupart des interférences existantes n'auraient pas d'effet notable sur l'interprétation du signal reçu. En nous basant sur les mesures effectuées sur le prototype expérimental, nous avons conclu qu'une puissance crête de l'ordre de 2 à 10 Watts serait suffisante. Plusieurs remarques sont faites dans le rapport au sujet des amplificateurs de puissance à semiconducteurs et au sujet des problèmes d'amplification des impulsions synthétiques.

3.1 Remarques à propos des amplificateurs de puissance

L'amplification du train d'impulsions synthétiques de l'émetteur peut se faire de deux manières:

- amplification directe du train d'impulsions synthétiques
- amplification séparée de chaque composante de fréquence contenue dans l'impulsion synthétique

Dans la méthode à amplification directe, on amplifie l'impulsion synthétique aussi linéairement que possible dans un amplificateur de puissance à transistors à large bande, fonctionnant en classe A. Le couplage à l'antenne nécessite alors l'utilisation d'un procédé d'isolation, de manière à éviter qu'une puissance réfléchie excessive en provenance de la charge n'atteigne l'étage de sortie. Deux cas sont possibles selon l'endroit où s'effectue la commutation.

- Si la commutation s'effectue entre la sortie de l'amplificateur et l'antenne, il est recommandé de disposer le commutateur de telle sorte que la puissance réfléchie de l'antenne soit absorbée avant de pouvoir atteindre l'étage de sortie de l'amplificateur.

- Si la commutation s'effectue avant l'amplificateur de puissance et si la désadaptation d'antenne est faible (TOS <2), on pourra coupler l'antenne directement, mais il sera nécessaire que l'amplificateur puisse transmettre le train d'impulsions commutées sans détériorations excessives des temps de montée et de descente.

Dans les deux cas, l'amplificateur aura les caractéristiques suivantes:

a) faible génération d'harmoniques
b) faible distortion d'amplitude et de phase
c) réponse transitoire nette
d) rendement acceptable

Un amplificateur de puissance linéaire, capable en principe de satisfaire aux spécifications ci-dessus, a été commandé chez "MICROWAVE POWER DEVICES INC." (type EWA-2015). Des essais effectués en liaison avec le prototype expérimental du système de télédétection ont montré que cet amplificateur pouvait fonctionner sur une gamme de fréquence de 20 MHz à 150 MHz avec une variation de gain de ±0,5 décibels. La puissance de sortie minimale en ondes entretenues est de 10 Watts pour une puissance d'entrée de 1mW. La réponse

transitoire est satisfaisante avec un temps de montée de 12 ns et un temps de descente de 25 ns. Le rendement est faible et seulement de l'ordre de 10%. La génération des fréquences harmoniques est telle que le second harmonique se trouve à -22 dB et le troisième à -12 dB par rapport à la fréquence fondamentale. Ce niveau de troisième harmonique n'est pas satisfaisant, car il va nuire au réglage de la forme de l'impulsion synthétique. L'impulsion synthétique est constituée de fréquences harmoniques dont les amplitudes et les phases sont indépendantes. Toute intermodulation excessive générée dans l'amplificateur de puissance va rentre difficile une commande indépendante des amplitudes et des phases. Il n'a pas été possible d'entreprendre une étude détaillée de l'influence des harmoniques sur la forme de l'impulsion synthétique, mais il est probable qu'un niveau de -30 dB à -40 dB sur tous les harmoniques soit nécessaire. Par conséquent, l'amplificateur de puissance disponible à l'heure actuelle ne semble pas convenir tout à fait à l'amplification de l'impulsion synthétique.

- Dans le mode d'amplification séparé, les problèmes d'intermodulation inhérents à la méthode précédente peuvent être éliminés en amplifiant chaque composante de fréquence dans un amplificateur séparé, à bande étroite, fonctionnant en classe C. Les signaux individuels sont alors ensuite additionnés linéairement en sortie. Les amplificateurs de puissance à transistors fonctionnant en classe C sont connus pour avoir un meilleur rendement que ceux fonctionnant en classe A; cependant la vitesse de commutation est réduite en raison de la bande passante étroite. Dans ces conditions, il est souhaitable d'effectuer la commutation entre l'amplificateur et l'antenne.

3.2 Objectifs futurs

Aucune décision précise n'a encore été prise à propos du mode d'amplification qui sera adopté lors des développement futurs du système de télédétection. Bien que les mesures d'intermodulation effectuées sur un amplificateur fonctionnant dans le mode direct n'aient pas été totalement satisfaisantes, nous pensons que ce mode d'amplification présente toutefois l'avantage de permettre la commutation avant l'amplification. Dans ces conditions, la fuite de puissance vers le récepteur sera plus faible que si des amplificateurs séparées étaient utilisés. De plus, les temps de montée et de descente de l'impulsion seront mieux préservés. Par conséquent, nous avons décidé de reconsidérer la possibilité d'utilisation du mode direct d'amplification lors de la phase de conception de la partie haute fréquence du prototype de développement.

4. COMMUTATEUR HAUTE FREQUENCE

Une unité d'aiguillage émission/réception comprend:

- un commutateur d'émission, servant à former le train d'impulsions synthétiques et à empêcher que la puissance réfléchie par les commutateurs et l'antenne ne puisse inter-agir avec l'amplificateur.

- un commutateur de réception, servant à empêcher la puissance d'émission d'atteindre le récepteur pendant toute la durée de l'impulsion émise.

Ces deux types de commutateurs doivent remplir les conditions suivantes:

a) isolation suffisante
b) faible perte d'insertion
c) tenue correcte en puissance
d) intermodulation faible, compatible avec une bonne préservation de l'impulsion synthétique
e) transitoires faibles de commutation
f) vitesse de commutation compatible avec l'altitude minimale de fonctionnement du système de télédétection.

Mise à part la tenue en puissance, le commutateur de réception est le plus critique. Par conséquent, seul ce commutateur a fait l'object d'une évaluation dans le cadre du présent contrat. Des résultats de mesures effectuées sur le commutateur de réception du prototype expérimental sont donnés dans le rapport, où sont discutées aussi d'autres structures de commutateurs susceptibles de convenir au prototype de développement.

4.1 Evaluation du commutateur

Une évaluation complète de ce commutateur a été faite dans le cadre de ce contrat. Les résultats montrent que:

- L'isolation du commutateur pour un signal d'entrée de 4 mW, dont la fréquence varie de 30 à 150 MHz, est d'au moins - 60 dB. Ce niveau est encore supérieur au niveau d'affaiblissement du signal utile du système et est tout juste acceptable.

- La perte d'insertion est de l'ordre de 4 dB pour un courant de 10 mA, ce qui est acceptable.

- La tenue en puissance du commutateur à l'état ouvert est limité à environ 15 mW. Ceci est dû au fait qu'aucune polarisation inverse des diodes Schottky n'est possible lorsque le commutateur est ouvert, par conséquent, le commutateur peut être facilement excité dans son état fermé lorsque le signal haute fréquence dépasse un certain niveau.

- Lorsque le commutation est fermé, les deuxième et troisième harmoniques sont en dessous de -30 dB pour un niveau d'entrée de 4 mW. Ce niveau est compatible avec une bonne préservation de la forme de l'impulsion synthétique.

- Les transitoires de commutation, bien que considérablement réduits par la symétrie du circuit, sont encore supérieurs à 20 mV lorsque l'impulsion d'excitation est de 100 mV. La durée des pointes transitoires est de 7 nanosecondes. Ce niveau présente un danger de surcharge du récepteur et donne naissance à une suroscillation inacceptable pendant l'intervalle de réception.

- La vitesse de commutation est de l'ordre de 20 ns.

4.2 Objectifs futurs

Selon les résultats précédents, ce commutateur ne présente pas la tenue en puissance exigée par le système futur de télédétection; de plus, les pointes transitoires de commutation sont encore trop importantes. Bien que le niveau des pointes transitoires ait été diminué grâce à un arrangement symétrique décrit dans le rapport, la tenue en puissance nécessite d'être améliorée. Il n'est pas possible d'effectuer des modifications sur le commutateur actuel car les transformateurs et les diodes de commutation sont inaccessibles. Par conséquent, nous avons décidé de revoir complètement la conception de l'unité d'aiguillage lors de la phase de développement futur. Des remarques sont faites dans le rapport à propos de la conception d'un nouveau commutateur.

TABLE OF CONTENTS

1.	INTRODUCTION		1
	1.1	General	1
	1.2	Report Content	1
	1.3	Reference Documents	1
2.	SAMPLE AND HOLD UNIT		2
	2.1	General	2
	2.2	Computation of Sampling Unit Frequency Response	4
	2.3	Sampling Modes	8
	2.3.1	Sampling of Aperiodic Signals	8
	2.3.2	Sampling of Periodic Signals	10
	2.4	Application to the Sea-Ice Thickness Sensor	11
	2.4.1	Constraints	12
	2.5	Future Objectives	22
3.	POWER AMPLIFIER		23
	3.1	General	23
	3.2	Power Amplifier Design Considerations	23
	3.2.1	Direct Amplification of the Synthetic Pulse	24
	3.2.2	Separate Amplification of Individual Frequency Components of the Synthetic Pulse	25
4.	RF SWITCH		27
	4.1	General	27
	4.2	Evaluation of the Experimental Prototype Receiver Switch	28
	4.3	New Switch Design Concepts	31
5.	RF SECTION PRELIMINARY ORGANIZATION		33
	5.1	General	33
	5.2	Elimination of the Even-numbered Frequencies from the Synthetic Pulse	33
	5.3	Elimination of the Variable Capacitance Diode Phase Shifters	36
6.	CONCLUSIONS		38
APPENDIX A	Measurements of a Sample and Hold Unit		A-1
APPENDIX B	Measurements of a Power Amplifier		B-1
APPENDIX C	Measurements of an RF Switch		C-1

LIST OF FIGURES

Figure 2.1	Block Diagram of Sample and Hold Unit	2
Figure 2.2(A)	Ideal Behaviour of Sample and Hold Unit	3
Figure 2.2(B)	Real Behaviour of Sample and Hold Unit	3
Figure 2.3	Equivalent Circuit of Sampling Unit	4
Figure 2.4	Sampling of a Sinewave	5
Figure 2.5	Variation of Hold Level with Frequency for the "Track and Hold" Mode	6
Figure 2.6	Variation of Hold Level With Frequency for the "True Sampling" Mode	7
Figure 2.7(A)	Spectrum of Aperiodic Signal	8
Figure 2.7(B)	Spectrum of Sampled Aperiodic Signal	9
Figure 2.8	Sampling Time for Aperiodic Signals	10
Figure 2.9	Sampling of Periodic Signals	10
Figure 2.10	Frequency Spectrum of Synthetic Pulse Train	12
Figure 2.11(A)	Time and Frequency Functions of Original Synthetic Pulse Train	15
Figure 2.11(B)	Time and Frequency Functions of Stretched Synthetic Pulse Train	15
Figure 2.12	Time Stretching Circuit Configuration	16
Figure 3.1	Broadband Class A Power Amplifier Followed by Absorptive Switch	24
Figure 3.2	Separate Class C Power Amplifiers Followed by Power Combiner and Absorptive Switch	26
Figure 4.1	T-R Switch Functional Arrangement	27
Figure 4.2	Circuit Diagram of Double Balanced Mixer "Mini-Circuits ZMY-3" Operated as a Switch	28
Figure 4.3	"Mini-Circuits ZMY-3" Switch in the Off State	29
Figure 4.4	"Mini-Circuits ZMY-3" Switch in the On State	30
Figure 4.5	New Balanced Switch Configuration Using PIN Diodes	31
Figure 5.1	Formation of Symmetrical Synthetic Pulse Train with no Even-Numbered Frequencies	34
Figure 5.3	Phase Shifting Through Variable Attenuation	36

1. INTRODUCTION

1.1 General

This report presents the work carried out at MPB Technologies Inc. on contract 0SD80-00128 between March 15, 1981 and October 25, 1981.

The first objective of the contract was to investigate the feasibility of the most critical elements of the sea-ice thickness sensor RF*section which will be needed for the development of a fully automated sensor. Prior to the start of this contract, these elements have been identified as being respectively:

- The sample and hold unit used for digitizing the received signal
- The RF power amplifier needed for boosting up the transmitted power
- The RF switches used in the transmit-receive unit to prevent the transmitted power from leaking directly into the receiver.

The second objective of the contract was to propose a preliminary configuration for the RF section of the sea-ice thickness sensor.

1.2 Report Contents

The basic requirements leading to the selection of specific configurations for the sample and hold unit, the RF power amplifiers and the RF switches are presented respectively in parts 2, 3 and 4 with some desigr. considerations and a description of these elements. The test set-ups used to evaluate the performance of these elements and the measurement results are presented respectively in appendices A, B and C. Then, important design considerations for the RF sections of the sea-ice thickness sensor developmental prototype are presented in part 5.

1.3 Reference Documents

The following documents provide a background to the work reported here:

a) "Development of a Remote Sea-Ice Thickness Sensor, Phase I, Stage 2a, Part I".
Final Report, Electronics Design MPB Technologies Inc. Report TP1573/1574, June, 1978, J.M. Keelty, J. Brana.

b) "Evaluation of Options for the Processing of Sea-Ice Thickness Radar Measurements", MPB Technologies Inc. TP 2857
J.M. Keelty, J. Brana.

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* Radio Frequency

2. SAMPLE AND HOLD UNIT

2.1 General

A sample and hold unit is normally used to convert an analog signal into a sequence of constant levels of the same duration. These levels are then converted into a bit sequence by an analog-to-digital converter so that the signal can be stored and processed digitally. The block diagram of a sample and hold unit is shown in Figure 2.1.

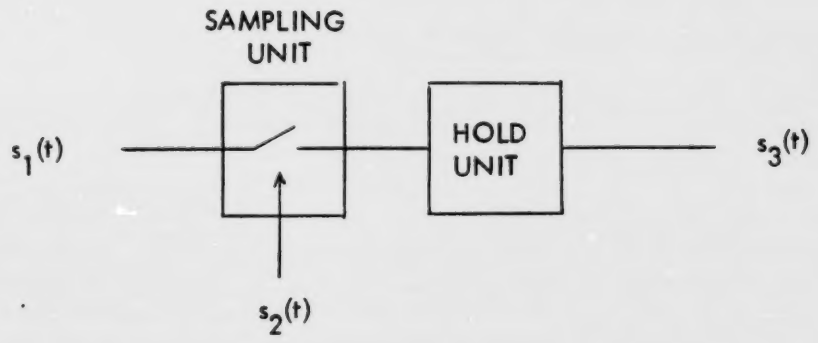


Fig. 2.1: Block Diagram of Sample and Hold Unit

The sampling unit consists of an electronic switch which is periodically closed by the gating signal $s_2(t)$. The input signal $s_1(t)$ is then transmitted through the switch during a short time called the sampling interval. The hold unit consists of an amplifier capable of memorizing for some time the value taken by the input signal $s_1(t)$ at the end of the sampling interval.

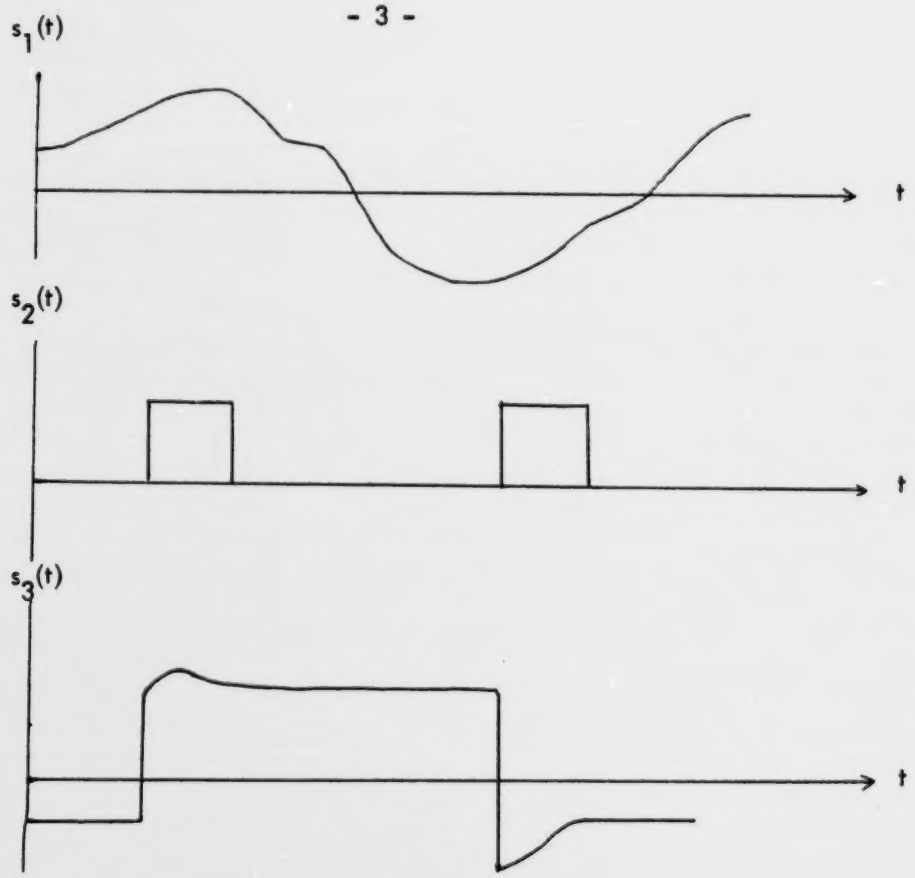


Figure 2.2(A): Ideal Behaviour of Sample and Hold Unit

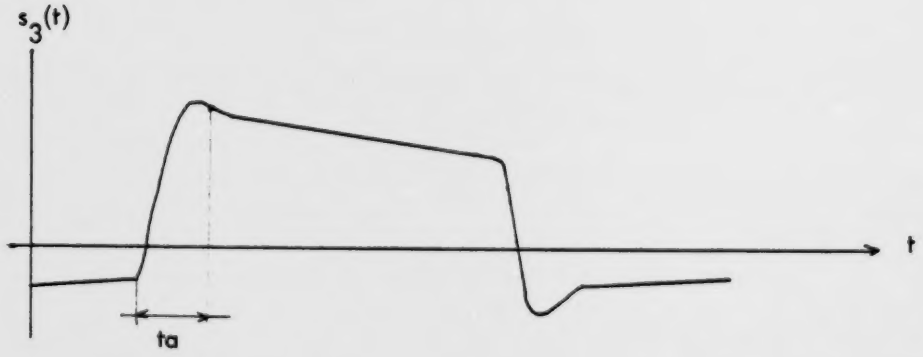


Figure 2.2(B): Real Behaviour of Sample and Hold Unit

In a perfect sample and hold unit, the time relationship between $s_1(t)$, $s_2(t)$ and $s_3(t)$ would be as shown in Figure 2.2(A). In this case, the output signal $s_3(t)$ starts to track the input signal $s_1(t)$ as soon as the switch closes at time t_1. Furthermore, the value of the input $s_1(t)$ at the end of the sampling interval is maintained perfectly constant until the next sample is collected. In a real sample and hold unit, the output signal $s_3(t)$ would look as shown in Figure 2.2(B). In this case, the output signal $s_3(t)$ only starts to track the input signal $s_1(t)$ after a time t_a usually called the acquisition time. Furthermore, the value of the input signal $s_1(t)$ at the end of the sampling interval is no more maintained constant until the next sample is collected, but decays by a certain amount called the percentage droop. Another parameter which is very important in practical situations is the 3 dB frequency bandwidth of the sample and hold unit. Acquisition time and 3 dB frequency bandwidth are interrelated parameters which give an indication of the sample and hold unit capability of handling fast varying signals.

2.2 Computation of Sampling Unit Frequency Response

The input signal $s_1(t)$ is generally applied to the sampling unit from a low resistance generator and the sampling unit is commonly terminated in the predominantly capacitive input impedance of the hold unit. Under these conditions, the sampling unit can be represented by the equivalent circuit of Figure 2.3.

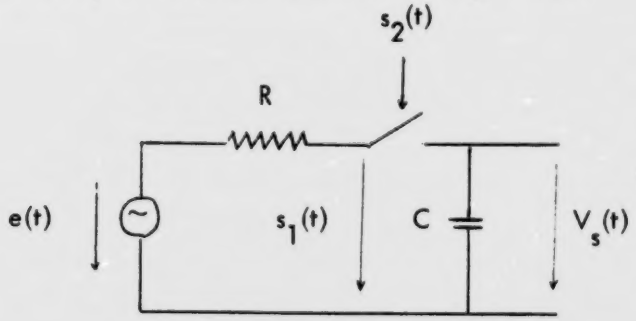


Figure 2.3: Equivalent Circuit of Sampling Unit

where:

R = generator internal resistance
C = hold unit input capacitance

In order to compute the frequency response of the sampling unit, the input signal will be assumed sinusoidal. Furthermore, capacitance C will be assumed completely discharged prior to sampling. Under these conditions, the hold level

at a given signal frequency can be determined from the following differential equation

$$\tau \frac{dv_s}{dt} + v_s = Ee^{j\omega t} \qquad (2.1)$$

$$\tau = RC$$

Therefore, after solving (2.1) over the sampling interval $t_2 - t_1 = \Delta t$, as shown in Figure 2.4,

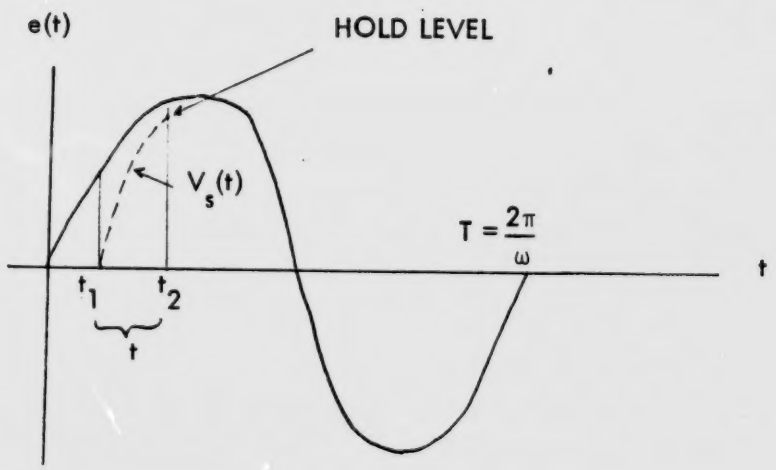


Figure 2.4: Sampling of a Sinewave

the hold level is found to be:

$$\overline{V}_s(t_2) = \left[\frac{1 - e^{-\frac{\Delta t}{\tau}} e^{-j\omega\Delta t}}{1 + j\omega\tau} \right] Ee^{-j\omega t_2} = \overline{V}_s e^{j\omega t_2} \qquad (2.2)$$

(the bar denotes complex quantities)

Now, the frequency response of the sampling unit is defined as:

$$H(j\omega) = \frac{\overline{V}_s}{E} = \frac{1 - e^{-\frac{\Delta t}{\tau}} e^{-j\omega\Delta t}}{1 + j\omega\tau} \tag{2.3}$$

the magnitude of $H(j\omega)$ is:

$$\left| H(j\omega) \right| = \sqrt{\frac{1 + e^{\frac{-2\Delta t}{\tau}} - 2e^{\frac{-\Delta t}{\tau}} \cos\omega\Delta t}{1 + (\omega\tau)^2}} \tag{2.4}$$

This last equation shows that the sampling unit can operate between two extreme modes of operation depending on the value of ratio $\frac{\Delta t}{\tau}$

- For $\frac{\Delta t}{\tau}$ very large, (2.4) degenerates into the following equation:

$$\left| H(j\omega) \right| = \frac{1}{\sqrt{1 + \left(\frac{\omega}{\omega_c}\right)^2}} \tag{2.5}$$

$$\omega_c = \frac{1}{\tau}$$

in this case, the hold level falls with frequency at a rate of 6 dB per octave as shown in Figure 2.5.

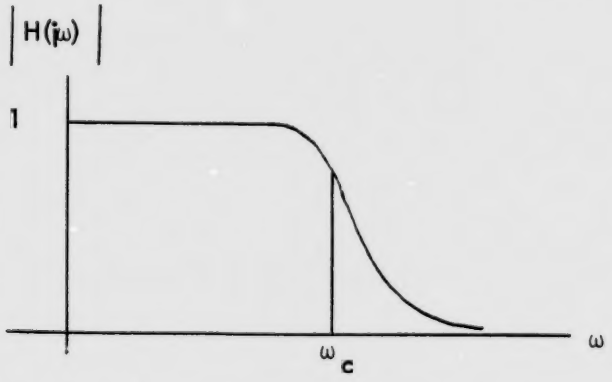


Figure 2.5: Variation of Hold Level With Frequency For the "Track and Hold" Mode

The 3 dB bandwidth or cutoff frequency of the sampling unit is thus simply:

$$f_c = \frac{\omega c}{2} = \frac{1}{2\pi\tau} \tag{2.6}$$

this mode of operation, where the sampling interval is much larger than the sampler time constant, is sometimes called "track and hold" mode.

- For $\frac{\Delta t}{\tau}$ very small, (2.4) degenerates into the following equation:

$$| H(j\omega) | = \left(\frac{\Delta t}{\tau}\right) \frac{\sin\left(\frac{\omega\Delta t}{2}\right)}{\left(\frac{\omega\Delta t}{2}\right)} \tag{2.7}$$

In this case, the hold level varies with frequency according to a damped oscillatory pattern, as shown in Figure 2.6.

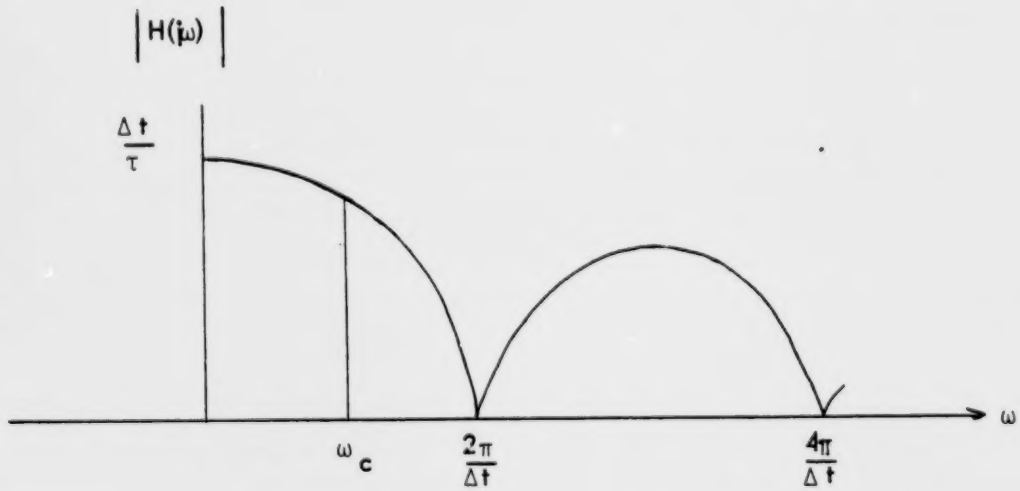


Figure 2.6: Variation of Hold Level with Frequency For the "True Sampling Mode"

The 3 dB bandwidth or cutoff frequency of the sampling unit is found to be independent of the time constant and given by

$$f_c = \frac{0.442}{\Delta t} \qquad (2.8)$$

Note that the d.c. gain is greatly reduced as compared with the previous case. This mode of operation, where the sampling interval is much smaller than the sampler time constant, is often called "true sampling" mode.

The choice between these two modes of operation depends on the signal-to-noise ratio at the sampling unit input, the required sampling frequency and the bandwidth requirements of the application at hand.

2.3 Sampling Modes

There are basically two modes of sampling -

- sampling of aperiodic signals
- sampling of periodic signals

These two modes will now be discussed below.

2.3.1 Sampling of aperiodic signals

Aperiodic signals are characterized by a continuous frequency spectrum as shown in Figure 2.7(A)

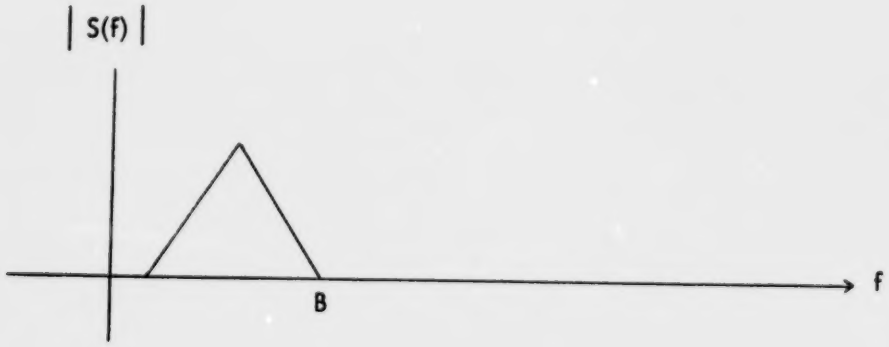


Figure 2.7(A): Spectrum of Aperiodic Signal

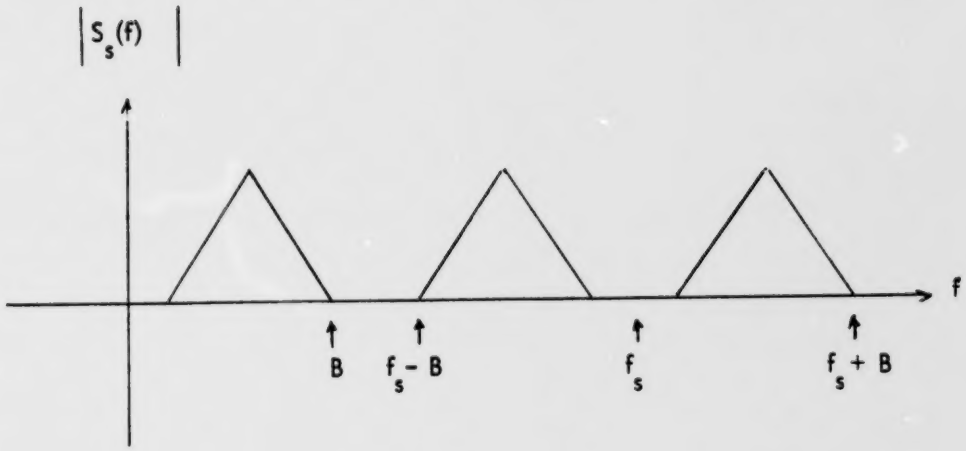


Figure 2.7(B): Spectrum of Sampled Aperiodic Signal

When such a signal is sampled at frequency f_s, the frequency spectrum of the sampled signal appears as shown in Figure 2.7(B). The spectrum of the original signal repeats itself around frequencies f_s, $2f_s$, ... Now, if the sampled signal is to retain the information of the original signal, the sampling frequency f_s must be chosen so that the various lobes do not overlap. From Figure 2.7(B), it appears that no overlap will occur if the sampling frequency satisfies the following inequality

$$f_s \geq 2B \tag{2.9}$$

where B is the bandwidth of the original signal (assumed to be band-limited for simplicity). Whenever (2.9) is satisfied, the original signal can be recovered through low pass filtering or through digital filtering of the sampled signal. Digital processing of the original signal requires that each sample be converted into a binary number before the next sample is taken. Consequently, the sampling time Δt must be made as short as possible and smaller than the sampling period T_s, as shown in Figure 2.8.

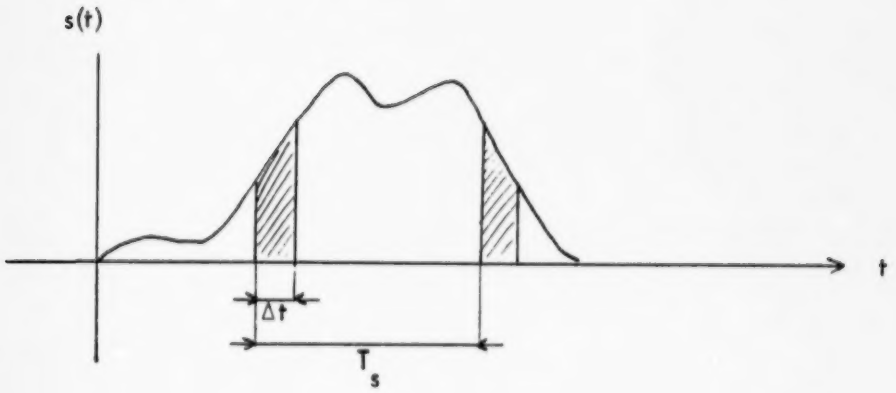


Figure 2.8: Sampling Time for Aperiodic Signals

In practical situations, T_s is limited by the speed capability of the analog-to-digital converter.

2.3.2 Sampling of periodic signals

When the signal is periodic of period T, it is not necessary to collect all the required samples during one period only. Instead, one sample can be collected at each period, as shown in Figure 2.9.

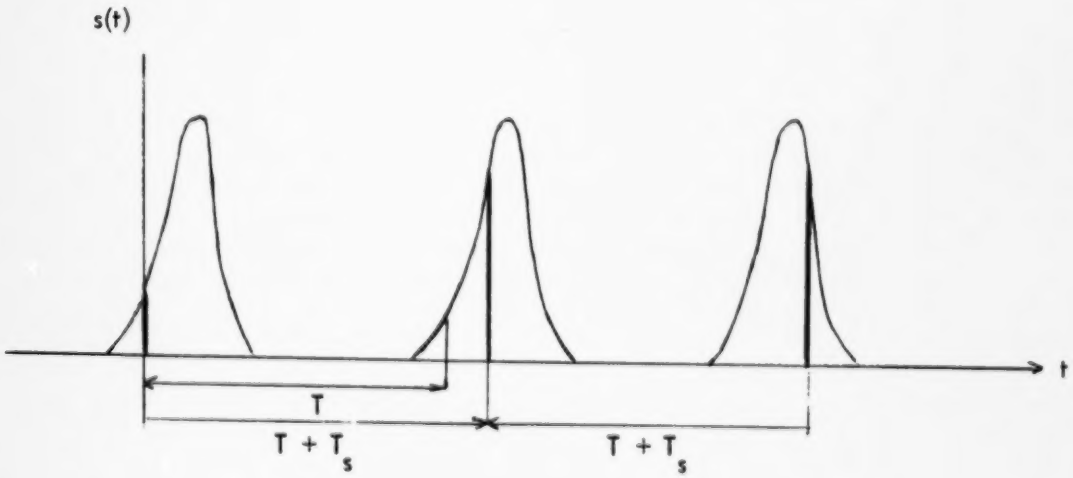


Figure 2.9: Sampling of Periodic Signals

If the sampling period is chosen to be $T_s' = T + T_s$, with $T_s = \frac{T}{N}$, the following sequence is generated:

$$\begin{aligned} s'(o) &= s(o) \\ s'(T + T_s) &= s(T_s) \\ s'(2T + 2T_s) &= s(2T_s) \\ &\text{----------} \\ &\text{----------} \\ s'\left[(N-1)T + (N-1)T_s\right] &= s(T-T_s) \end{aligned}$$

This sequence of samples could also have been obtained by taking N samples of a new periodic signal s'(t), having the same shape as the original signal, but of period

$$T' = NT \tag{2.10}$$

The bandwidth of the new signal is reduced to

$$B' = \frac{B}{N} \tag{2.11}$$

and the new sampling frequency becomes:

$$f_s' = \frac{1}{T + T_s} = \frac{f_s}{N + 1} \tag{2.12}$$

Finally, if s'(t) is to be recovered from its sampled version, the following conditions must be met

$$f_s' \geq 2B'$$

or, from (2.11) and (2.12),

$$T_s = \frac{1}{f_s} \leq \frac{N}{N + 1} \cdot \frac{1}{2B} \tag{2.13}$$

In other words, (2.13) defines the sampling time increment necessary to retain all the information of the original signal s(t).

2.4 Application to the Sea-ice Thickness Sensor

The constraints to be placed on a sampling and hold unit for the sea-ice thickness sensor will be now established and four possible configurations will then be discussed.

2.4.1 Constraints

In the synthetic pulse mode using K harmonically related frequencies, the frequency spectrum of the received signal is made up of K lobes as shown in Figure 2.10.

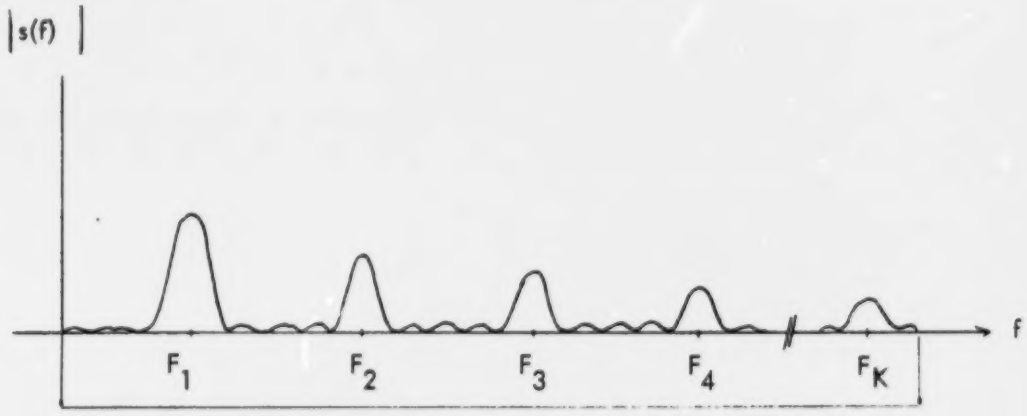


Figure 2.10: Frequency Spectrum of Synthetic Pulse Train

The frequency spectrum bandwidth is limited and given with a good approximation by:

$$B_s = F_1(K + \frac{2}{Q}) \qquad (2.14)$$

where Q is the number of synthetic pulses within the synthetic pulse train of duration $\frac{Q}{F_1}$.

Now, the sampling frequency and bandwidth constraints will be established in two cases, depending on whether the received signal can be considered aperiodic or periodic.

a) Aperiodic Case -

Strictly speaking, the received signal is aperiodic because the helicopter or the aircraft carrying the sensor is in constant motion and consequently the shape of the successive pulse trains is variable. In this case the data acquisition must take place within one pulse train duration.

i) Direct Aperiodic Sampling

If the received signal were to be sampled directly, the required sampling frequency would be, from (2.9) and 2.14)

$$f_{ss} \geq 2F_1(K + \frac{2}{Q}) \tag{2.15}$$

Similarly, to maintain frequency and phase dispersions as low as possible, the cut-off frequency of the sampling unit should be kept as high as possible. The choice $f_{cs} = 1.5\ f_s$ looks like a reasonable compromise and from section 2.2, the required sampling interval would be:

For the "track and hold" mode,

$$\Delta t_{ths} \leq \frac{5}{3\pi f_{ss}} \tag{2.16}$$

For the "true" sampling" mode,

$$\Delta t_{tss} \leq \frac{0.442}{1.5 f_{ss}} \tag{2.17}$$

In the present sensor experimental prototype, the synthetic pulse parameters are:

$$F_1 = 30 \text{ MHz}, \quad K = 5, \quad Q = 6$$

under these conditions,

$$f_{ss} \geq 2 \times 30 \times 10^6 (5 + \frac{2}{6})$$

$$f_{ss} \geq 320 \text{ MHz}$$

$$\Delta t_{ths} \leq \frac{5}{3 \times \pi \times 320 \times 10^6}$$

$$\Delta t_{ths} \leq 1.65 \text{ nanoseconds}$$

$$\Delta t_{tss} \leq \frac{0.442}{1.5 \times 320 \times 10^6}$$

$$\Delta t_{tss} \leq 0.92 \text{ nanoseconds}$$

Although these values could be achieved by extreme care in circuit layout, a sampling frequency larger than 320 MHz is too high to match the relatively lower speed of A/D converters and buffer memories available on the market. Consequently direct aperiodic sampling cannot be considered further.

ii) Stretched Pulse Aperiodic Sampling

One way to avoid the difficulties associated with direct sampling and still retain the concept of aperiodic sampling is to implement a time stretching circuit to convert the synthetic pulse train into a slower synthetic pulse train of smaller bandwidth. The time stretching process is illustrated in Figure 2.11(A) and 2.11(B). In Figure 2.11(A), the received pulse train s(t) is made up of Q synthetic pulses of duration $1/F_1$. Its frequency spectrum $|s(f)|$ is, as before, composed of K lobes centered on the component frequencies $F_1, \ldots, F_k$ of the synthetic pulse. The stretched pulse train and its frequency spectrum are shown in Figure 2.11(B). To achieve a reasonable preservation of the pulse train shape, it is necessary to maintain some separation between the spectra of each frequency component. This puts a constraint on the amount of bandwidth compression achievable in practice. In Figure 2.11(B), these spectra are assumed limited to one main lobe and two sidelobes. Therefore, under the best conditions, the minimum values of frequencies F_1', F_2' ....., F_k' are given by the following expressions:

$$F_1' = \frac{4}{Q} F_1$$

$$F_2' = 2F_1' \qquad (2.18)$$

$$F_K' = K F_1'$$

Consequently, the reduced bandwidth of the stretched pulse train becomes:

$$B = \frac{2F_1}{Q} (1 + 2K) \qquad (2.19)$$

One possible time stretching circuit configuration is represented in Figure 2.12.

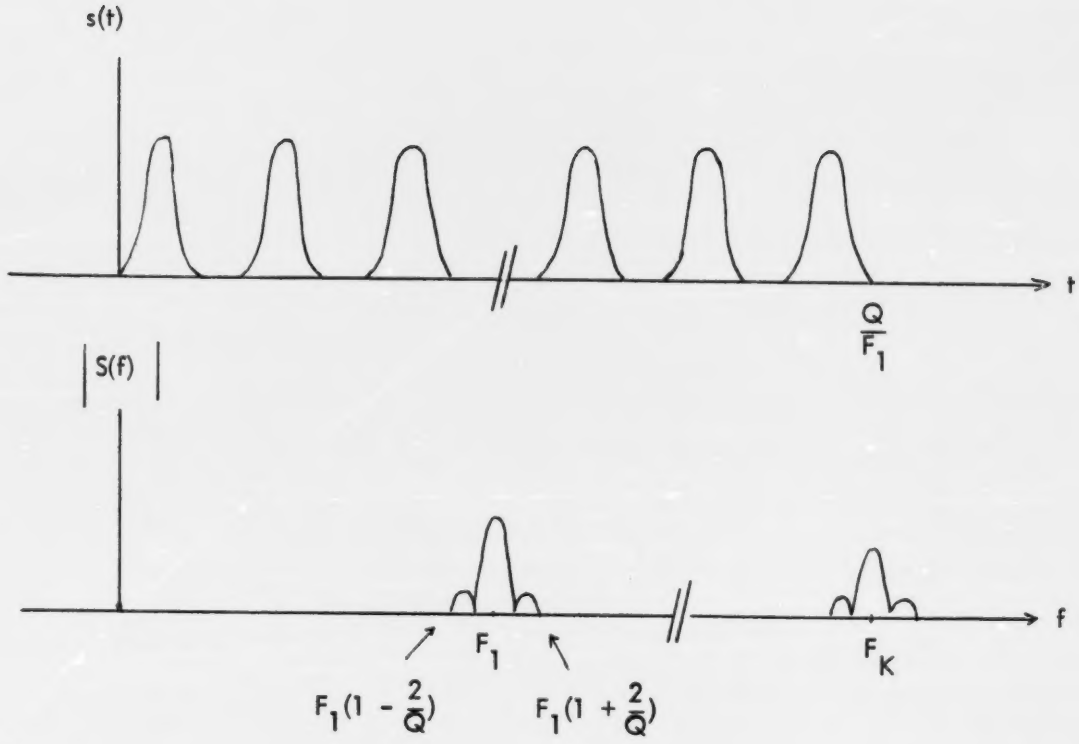


Figure 2.11(A): Time and Frequency Functions of Original Synthetic Pulse Train

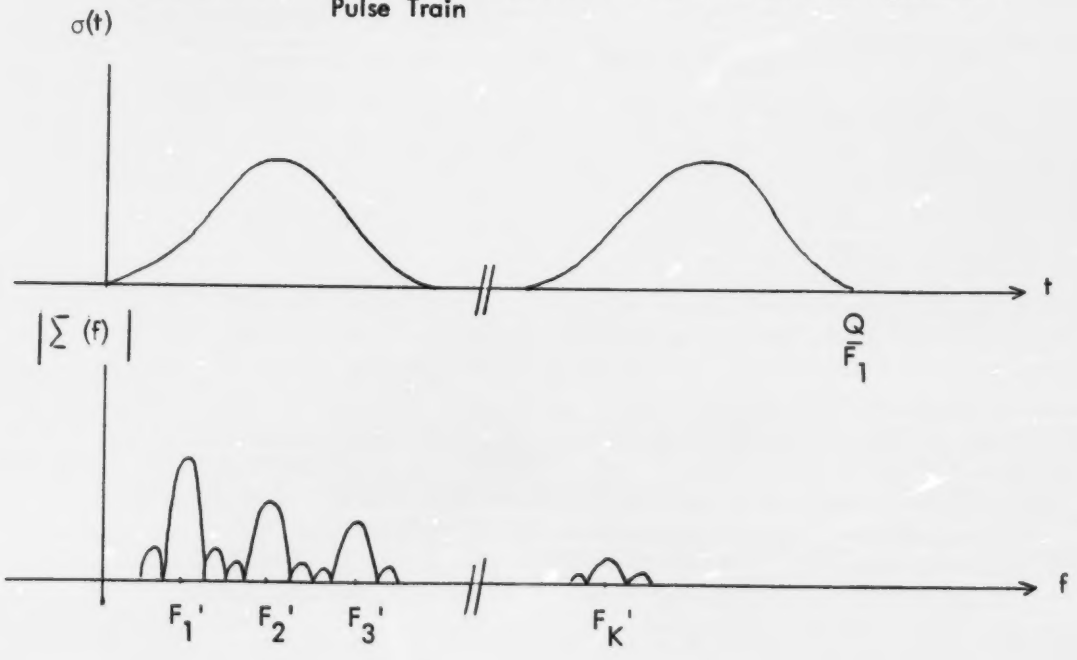


Figure 2.11(B): Time and Frequency Functions of Stretched Synthetic

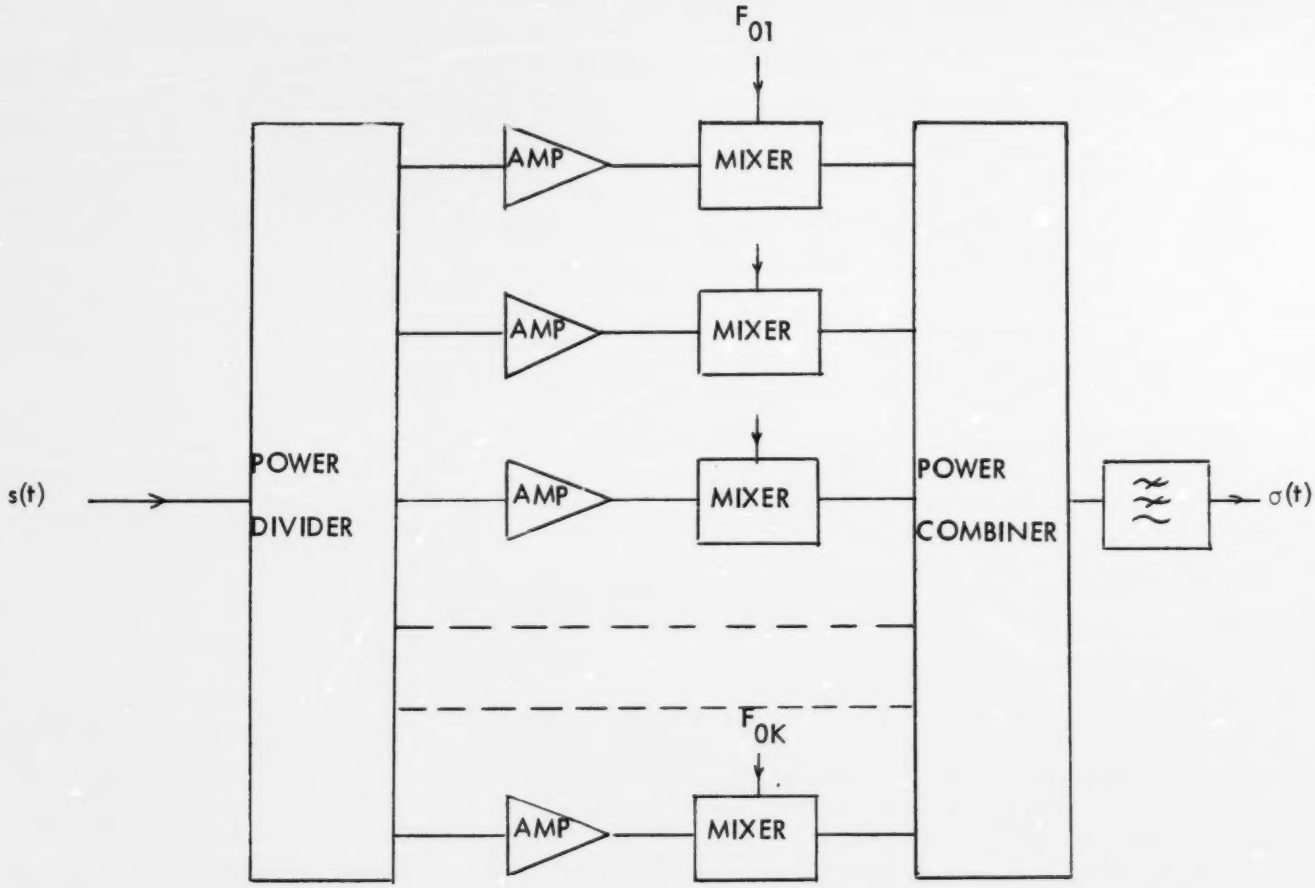


Figure 2.12: Time Stretching Circuit Configuration

The received synthetic pulse is applied to a broadband power divider and split into K identical synthetic pulse trains. These pulse trains are amplified and mixed individually with the following locally generated frequencies:

$$F_{01} = F_1(1 - \frac{4}{Q})$$

$$F_{02} = 2F_{01} \qquad (2.20)$$

$$F_{0K} = KF_{01}$$

The mixing products are then added linearly in a power combiner and applied to a low pass filter to produce four synthetic pulses of duration $\frac{Q}{4}F_1$

Now, coming back to the concept of aperiodic sampling, it is clear that the constraints on sampling parameters can be greatly reduced by using the pulse stretching approach described above. In particular, expressions (2.15), (2.16), and (2.17) are modified to become

$$f_{s\sigma} \geq \frac{4F_1}{Q}(1 + 2K) \qquad (2.21)$$

$$\Delta t_{th\sigma} \leq \frac{5}{3\pi f_{s\sigma}} \qquad (2.22)$$

$$\Delta t_{tss} \leq \frac{0.442}{1.5 f_{s\sigma}} \qquad (2.23)$$

Applications of these concepts to the sea-ice thickness sensor implies that the received synthetic pulse train should be stretched to the point where existing sample and hold units, A/D converters and PROM's could be used. A survey of several manufacturers shows that the combined speed of these devices is limited by the access time to the PROM's. For instance, one of the fastest 8 bit bipolar PROM , the "INTEL 3636", has a maximum access time of 65 ns and therefore allows a maximum sampling frequency of the order of f_s = 15 MHz. Thus, for F_1 = 30 MHz and K = 5, the value of Q must be, from (2.21)

$$Q \geq 4 \frac{F_1}{f_{s\sigma}} (1 + 2K)$$

$$Q \geq 4 \frac{30}{15} (1 + 2 \times 5)$$

$$\underline{Q \geq 88}$$

Also, from 2.22 and 2.23,

$$\Delta t_{th\sigma} \leq \frac{5}{3\pi \times 15 \times 10^6}$$

$$\underline{\Delta t_{th\sigma} \leq 35 \text{ nanoseconds}}$$

$$\Delta t_{ts\sigma} \leq \frac{0.442}{1.5 \times 15 \times 10^6}$$

$$\underline{\Delta t_{ts\sigma} \leq 20 \text{ nanoseconds}}$$

From (2.20), the reduced frequencies for Q = 100 would be respectively:

$F_1' = 1.2$ MHz

$F_2' = 2.4$ MHz

$F_3' = 3.6$ MHz

$F_4' = 4.8$ MHz

$F_5' = 6$ MHz

From (2.19), the stretched pulse train bandwidth would be:

$$B_\sigma = \frac{2 \times 30 \times 10^6}{100} (1 + 2 \times 5)$$

$$\underline{B_\sigma = 6.6 \text{ MHz}}$$

Similarly, from (2.20), the required local frequencies would be respectively:

F_{01} = 28.8 MHz

F_{02} = 57.6 MHz

F_{03} = 86.4 MHz

F_{04} = 115.2 MHz

F_{05} = 144.0 MHz

The following remarks can be made:

1) Aperiodic sampling of the stretched pulse train could considerably reduce the speed and bandwidth constraints of the sample and hold unit as well as the speed constraints of the A/D converter and buffer memories. A sample and hold unit capable of operating conservatively at a sampling frequency of 15 MHz and having a bandwidth of 50 MHz has been procured from "DATA DEVICES CORPORATION". This unit (type DDC-8530) has been fully evaluated at MPB Technologies Inc. The test set-ups and the results are described in details in Appendix A. The results show that this unit would introduce very small attenuation and phase dispersion over the 6.6 MHz bandwidth of the stretched pulse train.

2) Aperiodic sampling of the stretched pulse train offers the advantage of allowing the acquisition of all the required data during one repetition period of the sensor transmitted signal. Therefore the motions of the helicopter or the aircraft carrying the sensor will have a negligible effect on the received data. The disadvantage is that a large number of synthetic pulses are required and this puts a constraint on the altitude from which the sensor must operate to avoid interferences between the transmitted and received signals. In the above example, ~100 synthetic pulses are required in the original pulse train, leading to a pulse train duration of 3.3 The helicopter or the aircraft would then have to operate at an altitude of 470 metres and the power requirements would have to be increased accordingly. Note that if the two sidelobes in Figure 11B were neglected, the minimum operating altitude could be halved to 235 metres. However, in this case, the shape preservation of stretched synthetic pulses might not be as

satisfactory, but this has still to be demonstrated by computer simulation.

3) The number of samples to be collected depends on the pulse train repetition frequency. Thus, for a repetition frequency of 50 KHz, the number of samples to be collected would be

$$N = \frac{15 \times 10^6}{50 \times 10^3} = 300$$

This number is 320/15 ≃ 21 times smaller than what would have been required with direct aperiodic sampling. A single 1 Kbyte (8 bit) PROM of the INTEL 3628 type would thus be sufficient for this application.

4) Another advantage of this method of sampling is that no scanning time base is necessary. The sampling frequency could be easily synchronized with the sensor basic clock and the effect of time jitter would be negligible.

b) Periodic Case -

The direct sampling and stretched pulse sampling methods described above in the aperiodic case can be readily applied to the periodic case. However, periodic sampling can only be used if the shape of the successive pulse trains does not change appreciably over the time required to acquire all the necessary samples. The main advantage of periodic sampling is that the constraints on sampling speed, A/D conversion time and access time to buffer memories are considerably relieved. However, the bandwidth constraint for the sample and hold unit is maintained. The cases of direct periodic sampling and stretched pulse periodic sampling will now be considered.

i) Direct Periodic Sampling

Assume again for the synthetic pulse,

$$F_1 = 30 \text{ MHz}, \quad K = 5, \quad Q = 6$$

Then, for a 50 KHz repetition rate, the number of samples to be collected would be:

$$N = \frac{320 \times 10^6}{50 \times 10^3} = 6400$$

As only one sample is collected at each repetition period, the total

acquisition time is, from (2.10),

$$T' = 6400 \times \frac{10^{-3}}{50}$$

$$\underline{T' = 128 \text{ milliseconds}}$$

During that time, a helicopter moving horizontally at 50 km/h would only change its position by 2.5 metres and the error of interpretation is likely to be quite small. The required bandwidth for the sampling and hold unit is the same as in the direct aperiodic sampling case. In other words

$$\underline{f_{cs} = 1.5 f_s = 480 \text{ MHz}}$$

Similarly, the required sampling intervals are, for the "track and hold" mode and the "true sampling" mode respectively,

$$\underline{t_{ths} \quad 1.65 \text{ nanoseconds}}$$

$$\underline{t_{tss} \quad 0.92 \text{ nanoseconds}}$$

Also, from the considerations of subsection 2.3.2, the sampling time increment should be

$$T_s = \frac{T}{N} = \frac{1}{6400 \times 50 \times 10^3}$$

$$\underline{T_s = 3.125 \text{ nanoseconds}}$$

Direct periodic sampling is very attractive and is certainly the solution to adopt if the high frequency problems associated with the short sample interval and the scanner time jitter can be solved. These problems are under current investigation at MPB Technologies Inc. and fast Schottky diodes, required for the sampling unit, have been ordered. Assuming that these problems can be solved, only low speed A/D converters and buffer memories, operating at 50 KHz, would be required. It is also important to note that this approach would not put any altitude constraints on the helicopter.

ii) Stretched Pulse Periodic Sampling

In the case of periodic sampling, the advantages of the pulse stretching techniques described for the aperiodic case are partly lost because there is no more any need to lower the sampling frequency. The method would still be useful to reduce the bandwidth requirements of the sample and hold unit if the direct periodic sampling method were not satisfactory. However,

it must be noted that the resulting scheme would increase the minimum altitude at which the helicopter would have to operate. This increase would depend upon the desired bandwidth reduction.

2.5 Future Objectives

A comparison of the advantages and disadvantages of the four sampling techniques shows that the direct periodic sampling method is the best suited to the ice thickness sensor for the following reasons:

- no limitations of the minimum operating altitude of the sensor
- no frequency conversion required
- only low speed A/D converters are necessary
- critical high frequency operation limited to sampling gate and associated driver circuitry

It has consequently been decided that a direct periodic sample and hold unit and associated scanning circuitry will be developed during the RF section design phase of the sensor developmental prototype.

3. POWER AMPLIFIER

3.1 General

The peak power required at the sea-ice thickness sensor transmitter output is a function of the peak signal-to-noise ratio tolerable at the receiver input. It is commonly understood that a peak signal-to-noise ratio of 20 to 30 dB at the receiver will allow adequate resolution of the received signal. For a given noise level at the receiver input, the peak signal power required at the transmitter output will then be a function of the following parameters:

a) operating altitude
b) reflection coefficients for the upper and lower sea-ice boundaries
c) antenna gain characteristics

The helicopter operating altitude will likely not exceed 300 metres. The reflection coefficients for the ice upper and lower boundaries have known limits for the sea-ice thicknesses under consideration in this project.

The gain characteristics of the newly developed antenna are under current evaluation and will not be known accurately before some form of trial in the field takes place.

The antenna unknown characteristics, coupled to the fact that the noise level at the receiver input can vary within very large limits depending on where the sensor is operated, make it very difficult to predict any transmitter power requirements at the moment. In the arctic sea, the noise is predominantly thermal and generated in the antenna and receiver electronics. In other regions that might be selected to carry out preliminary trials, the electromagnetic interferences from nearby VHF transmitters become the predominant source of noise. Under these conditions, it was decided to increase the maximum peak power of the transmitter to such a level that even the worst interferences would not have any significant effects on the received signal interpretation. Based on previous measurements made on the sea-ice thickness sensor experimental prototype, it was felt that a transmitter peak power ranging from 2 to 10 watts would be adequate. Design considerations on solid state power amplifiers and some of the problems encountered when trying to amplify synthetic pulses will be discussed in the remainder of this section.

3.2 Power Amplifier Design Considerations

The amplification of the synthetic pulse train at the transmitter end can be achieved in two ways:

- Direct amplification of the synthetic pulse
- Separate amplification of indifidual frequency components of the synthetic pulse

These two methods will now be discussed.

3.2.1 Direct Amplification of the Synthetic Pulse

In this case, the synthetic pulse is amplified as linearly as possible through a broadband VHF power transistor amplifier operating in a class A mode. The connections to the antenna must be made through some form of isolating device to prevent excessive reflected power from the load to reach the output transistor stage. Two cases are possible depending on where switching takes place.

- If switching is done between the amplifier output and the antenna, the transmitter switch could be embedded in an absorptive type of arrangement as shown in Figure 3.1. In this configuration, the output signal of the amplifier is split between the two switches by means of a 0° - 90° 3 dB directional coupler which will send the power reflected from the two switches to the dummy load R_1 and prevent it from reaching the power amplifier output. Similarly, any mismatch at the antenna will send the reflected power back to the dummy load R_2.

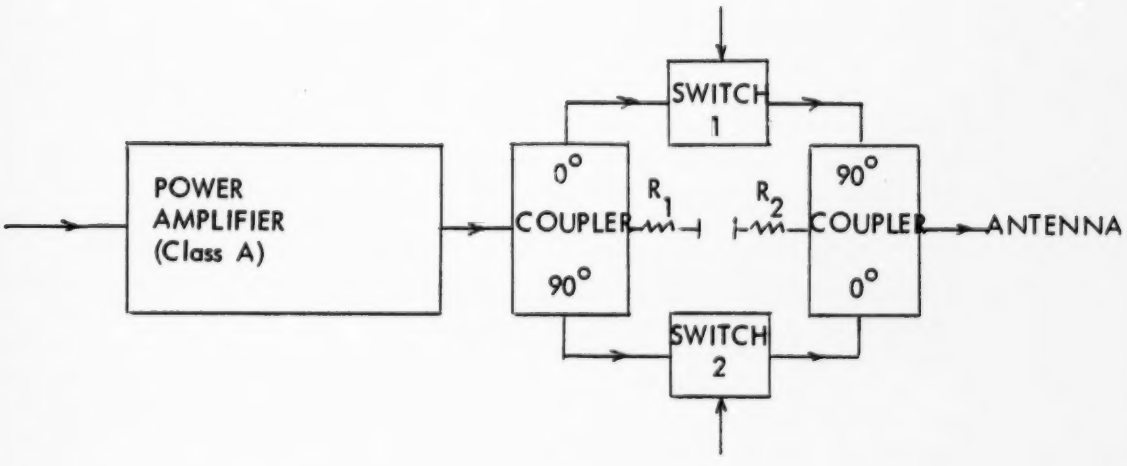


Figure 3.1: Broadband Class A Power Amplifier Followed by Absorptive Switch

- If switching is done before the power amplifies and if the antenna mismatch is kept small (VSWR < 2), the antenna can be connected directly, but the amplifier must be capable of transmitting the switched pulse train without significant rise time and fall time degradations.

In both cases, the power amplifier must have the following characteristics:

a) low harmonic generation
b) low amplitude and phase dispersion with frequency
c) clear transient response
d) reasonable efficiency

A linear power amplifier presumably supposed to meet these requirements has been procured from "MICROWAVE POWER DEVICES INC." (type EWA-2015) and tested in conjunction with the sea-ice thickness sensor experimental prototype. Details on the measurement procedures and results are given in Appendix B. The amplifier is capable of operating over the frequency range 20 MHz to 150 MHz with a ±0.5 dB gain variation. The amplifier nominal CW output power is 10W for an input power of 1mW. The transient response is satisfactory with a 12 ns rise time and a 25 ns fall time. The efficiency is low and of the order of 10%. The harmonic generation was found to be -22 dB for the second harmonic and only -12 dB for the third harmonic. This level of third harmonic generation is not satisfactory, because it will impair the adjustment of the synthetic pulse shape. The synthetic pulse is made up of harmonically related frequency components with independent amplitude and phases. Excessive intermodulation within the power amplifier will make independent amplitude and phase control impossible. A detailed study of the influence of harmonic generation on the synthetic pulse shape has not been undertaken yet, but it is likely that a level of -30 dB to -40 dB on all harmonics will be necessary. Therefore, the power amplifier presently available does not appear suitable for synthetic pulse amplifications.

3.2.2 Separate Amplification of Individual Frequency Components of the Synthetic Pulse

The intermodulation problems encountered in the previous method of amplification can be eliminated by amplifying each frequency component in separate narrowband class C amplifiers. The individual signals are then linearly added at the output. Transistor class C power amplifiers have also a better efficiency than class A amplifiers but the switching speed is affected by the narrower bandwidth. Under these conditions, switching will have to be performed between the power amplifier and the antenna, as shown in Figure 3.2.

The switch arrangement is absorptive and the same as in Figure 3.1. The power combiner could be made of several branching filters tuned at frequencies F_1, F_2,... F_K or could be a straight K-way broadband power combiner with appropriate power rating. The individual power amplifiers would be much simpler than the broadband unit considered earlier, because there is no need to trade gain for bandwidth in this case. Although no practical implementation of this scheme has been made, further investigation into this concept will be carried out during the design phase of the developmental prototype RF section.

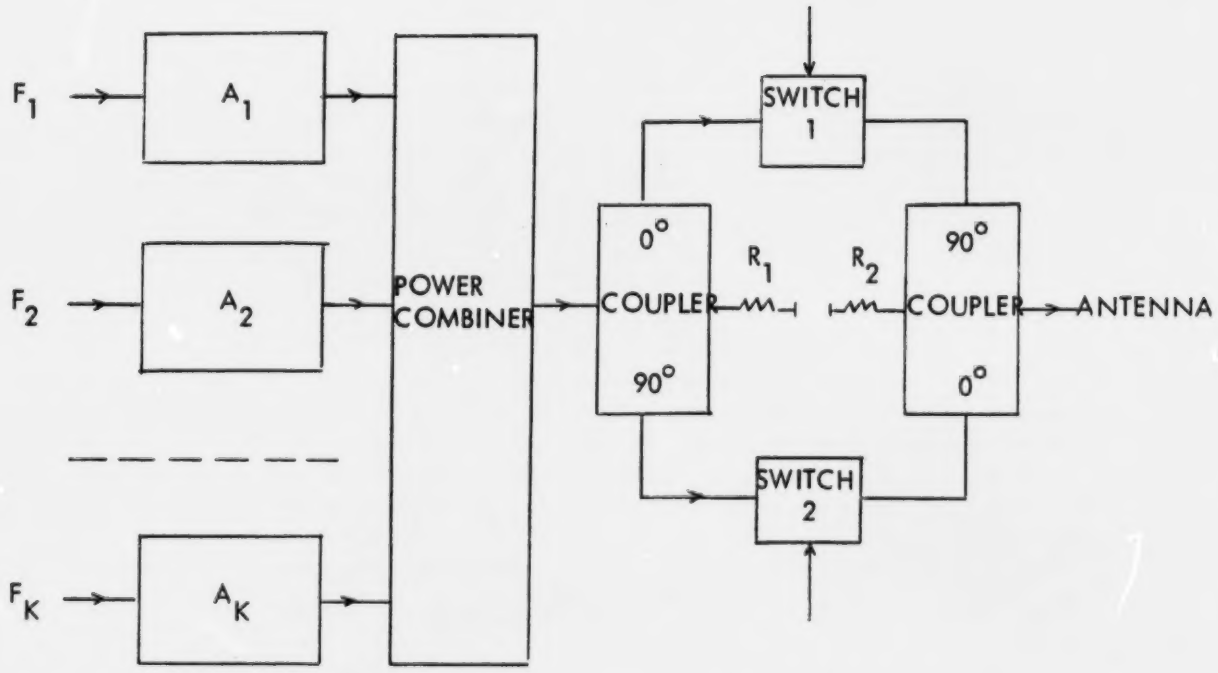


Figure 3.2: Separate Class C Power Amplifiers Followed by Power Combiner and Absorptive Switch

4. RF SWITCH

4.1 General

A basic transmit-receive (T-R) switch configuration suitable for a sea-ice thickness sensor operating in the synthetic pulse mode is shown in Figure 4.1.

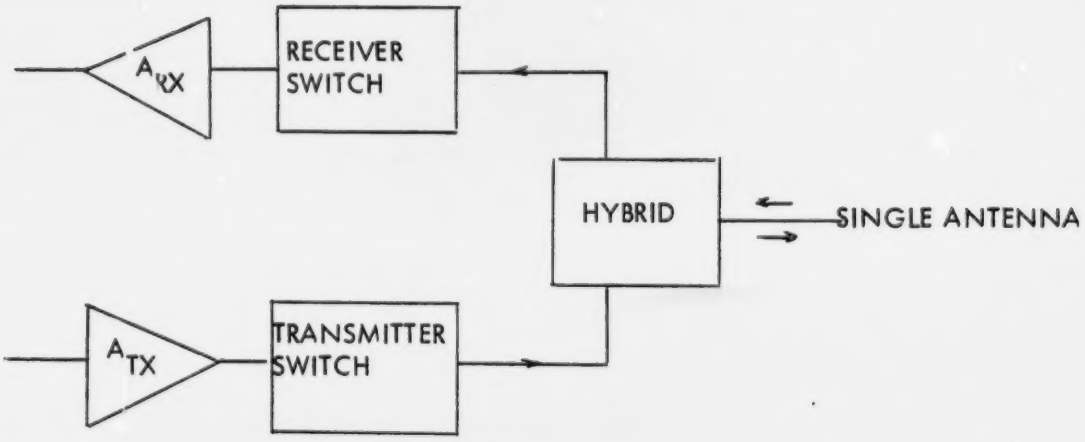


Figure 4.1: T-R Switch Functional Arrangement

In this configuration,

- The transmitter switch is used to gate the synthetic pulse train and prevent excessive reflected power from the switching devices and the antenna to be sent back to the power amplifier.
- The receiver switch is used to prevent the transmitter from leaking into the receiver during transmit time.

Both switches must meet the following general constraints:

a) sufficient isolation
b) low insertion loss
c) suitable power handling capability
d) low intermodulation, compatible with synthetic pulse shape preservation
e) low switching transients
f) switching speed compatible with the minimum operating altitude of the sensor.

Except for the power handling capability, the receiver switch is more critical than the transmitter switch. Consequently, only the receiver switch has been investigated within the frame of the present contract. Measurements made on the receiver switch used in the sea-ice thickness sensor experimental prototype are reported here and alternative switch configurations suitable for the developmental prototype are discussed.

4.2 Evaluation of the Experimental Prototype Receiver switch

This switch was procured from "MINI-CIRCUITS" (type ZMY-3) and consists of a Schottky diode double balanced mixer operated as a switch. The circuit diagram is represented in Figure 4.2.

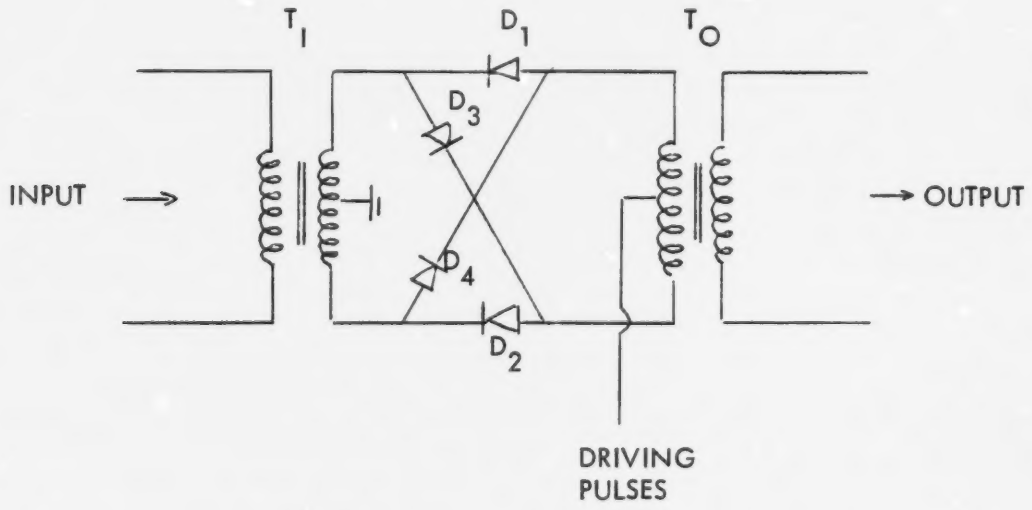


Figure 4.2: Circuit Diagram of Double Balanced Mixer "Mini-Circuits ZMY-3" Operated as a Switch

The signal received from the antenna is applied to a Schottky diode matched quad through the centertapped transformer T_I. A driving pulse is applied to the centertap of transformer T_O.

In absence of a driving pulse, each diode of the quad is at zero potential and the switch is in its off state. Under these conditions, the diodes look capacitive and the off-state equivalent circuit appears as in Figure 4.3.

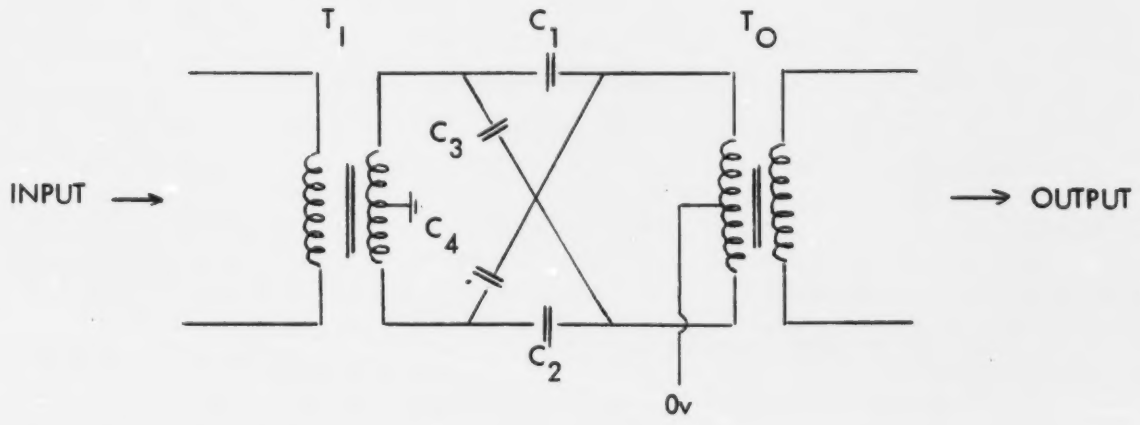


Figure 4.3: "Mini-Circuits ZMY-3" Switch in the Off-State

Capacitances C_1 to C_4 are small, presumably equal, and form a balanced bridge. The voltages appearing at terminals a and b of T_O are equal and out of phase. Thus, they cancel in transformer T_O and the switch theoretically offers an infinite isolation. In practice, the isolation cannot be made infinite because the transformers are never perfectly symmetrical and the diodes are never exactly identical.

In presence of a positive pulse, diodes D_1 and D_2 are forward-based, whereas diodes D_3 and D_4 are strongly back-biased. The switch is thus in its on state. The on state equivalent circuit appears as in Figure 4.4.

Resistances R_1 and R_2 are relatively small compared with the source and load resistances and consequently the switch insertion loss is small. It must also be noted that the switch isolates the driving pulse from the input and output, but the degree of isolation again depends on the circuit symmetry.

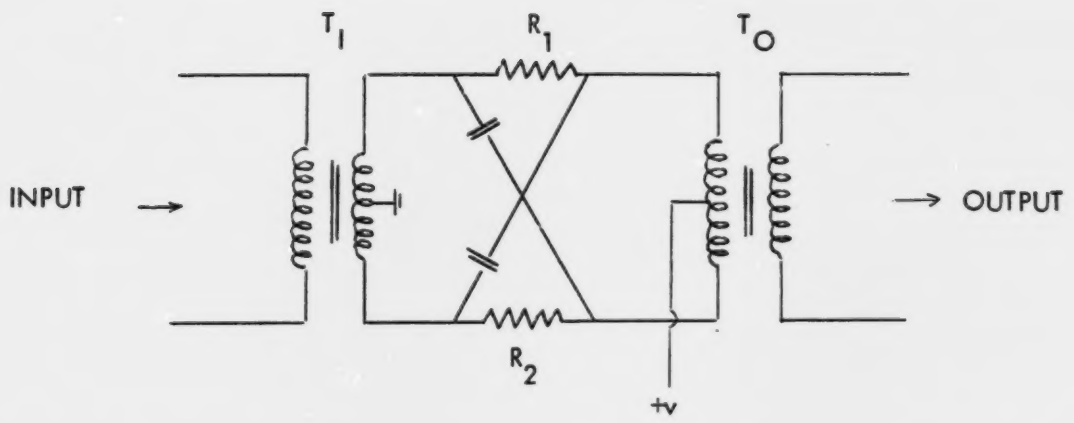


Figure 4.4: "Mini-Circuits ZMY-3" Switch in the On-State

A complete evaluation of this switch was made within the frame of this contract. The test set-ups and the measurement results are presented in Appendix C. The following comments can be made:

- The switch isolation for a 4 mW input signal varying in frequency from 30 MHz to 150 MHz is at least -60 dB. This is still above the level of the expected attenuation of the sensor signal path and is just acceptable.
- The insertion loss is about 4 dB for a 10 mA driving current. This is acceptable.
- The power handling capability of the switch in the off state is limited to approximately 15 mW. This is due to the fact that no back-biasing of the diodes is possible when the switch is in the off state and therefore the switch can be easily driven into the on state when the RF signal exceeds a certain power level.
- The second and third harmonic generation in the on state are below -30 dB for a 4 mW input power level. This is compatible with good preservation of the synthetic pulse shape.
- The switching transients, although already reduced by the circuit symmetry, and in excess of 20 mV for a 100 mV driving pulse accross the switches. The spikes duration is of the order of 7 nanoseconds. This level of transients has been found to overload the receiver and create unacceptable ringing during receiving time.

- The switching speed is of the order of 20 ns.

The previous results lead to the conclusions that the switch does not have the power handling capability required for the developmental prototype and that the switching transients are still too strong. Although the transient level has been reduced by the symmetric arrangement described in Appendix C, the power handling capability still needs improvements. No modifications can be made to the present switch, because the transformers and the diodes are not accessible. Therefore, it has been decided to review the design of the T-R switch for the developmental prototype. Although no realization has been made yet, design considerations for a new switch are presented below.

4.3 New Switch Design Concepts

The main difficulties encountered with the previous receiver switch are limited power handling capability and too high switching transients. The power handling capability can be increased by using PIN diodes instead of Schottky diodes. PIN diodes are a little slower than Schottky diodes, but they can still meet the speed requirements of the sensor. Furthermore, PIN diodes introduce much lower intermodulation and have a lower insertion loss than Schottky diodes. The balanced switch configuration of Figure 4.5 offers the possibility of achieving efficient diode back-biasing and transient control.

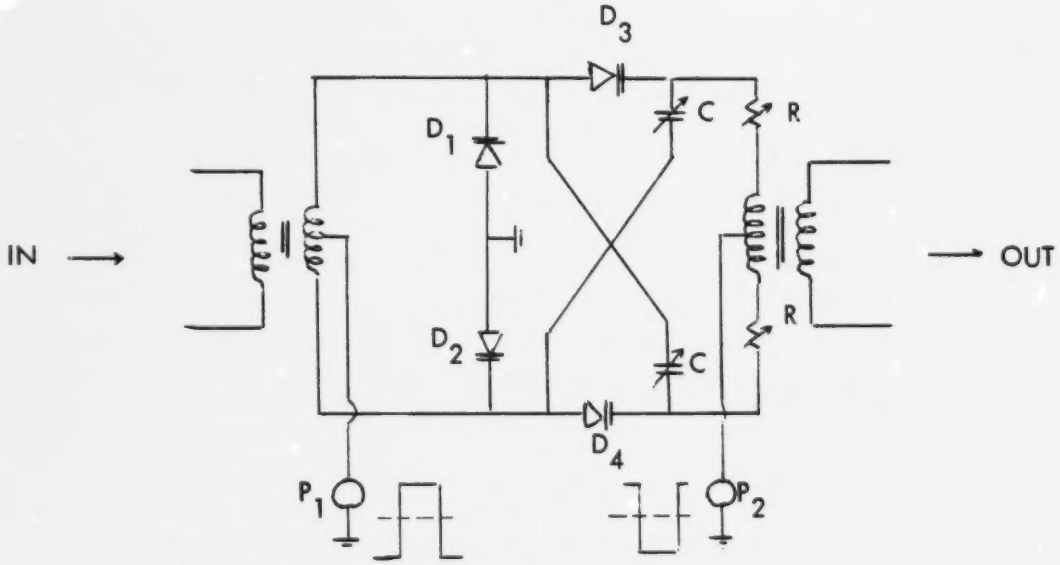


Figure 4.5: New Balanced Switch Configuration Using PIN Diodes

In this circuit, PIN diodes D_1, D_2 operate in the shunt mode and D_3, D_4 in the series mode. The switch is activated by two pulses of opposite polarities applied to the centertap of each transformer. In the off state, diodes D_1, D_2 are forward-biased and D_3, D_4 strongly back-biased. In the on state, the opposite situation occurs. The power handling capability of the switch in the off state depends on the back-biased voltage of diodes D_3, D_4 and is ultimately limited by their breakdown voltage. The switching transients appearing at the output can be controlled by the balancing capacitors C and resistors R. Further investigation of the particular switch will be made during the design phase of the RF section developmental prototype.

5. RF SECTION PRELIMINARY ORGANIZATION

5.1 General

It has not been possible within the frame of this contract to elaborate preliminary circuit diagrams for the developmental prototype RF section. The main reasons are the following:

a) The new antenna characteristics (input standing wave ratio and overall transfer function for the five synthetic pulse frequencies) were not available. These characteristics must be known fairly accurately before any gain and phase shift compensation scheme can be designed. They are also important for the determination of the transmitter power and receive switch isolation.

b) The synthetic pulse frequencies of 30 MHz, 60 MHz, 90 MHz, 120 MHz and 150 MHz are hardly compatible with the "COLLINS" AR-201 transceiver installed in Canadian Coast Guard helicopters and operating at 156 MHz. The switching of the upper frequency (150 MHz) would generate unacceptable spectral components within the passband of the transceiver. Under these conditions, the helicopter crew would not be able to communicate or exchange data with the ship while the sensor is operating. There are basically two methods of solving this problem, either modify the synthetic pulse frequencies or use a UHF transceiver in the helicopter. The suppression of the 150 MHz component is one possibility, but it will decrease the synthetic pulse sharpness. The extent of this pulse deterioration has not yet been investigated.

c) Unexpected difficulties were encountered during the testing of the sample and hold units, the power amplifier and the receiver RF switch. Consequently, more insight into these components was necessary.

However, two concepts have been elaborated which would considerably simplify the hardware of the RF section.

- Elimination of the even-numbered frequencies from the synthetic pulse
- Elimination of the variable capacitance diode phase shifters

These two concepts will now be briefly discussed.

5.2 Elimination of the Even-numbered Frequencies from the Synthetic Pulse

It is believed that the even frequency components could be eliminated from the synthetic pulse without affecting the resolution capability of the sensor. This can be explained by examining the waveforms of Figure 5.1.

Figure 5.1(A) shows the basic train of synthetic pulses of period T. Figure 5.1(B) shows the same train inverted and delayed by T/2.

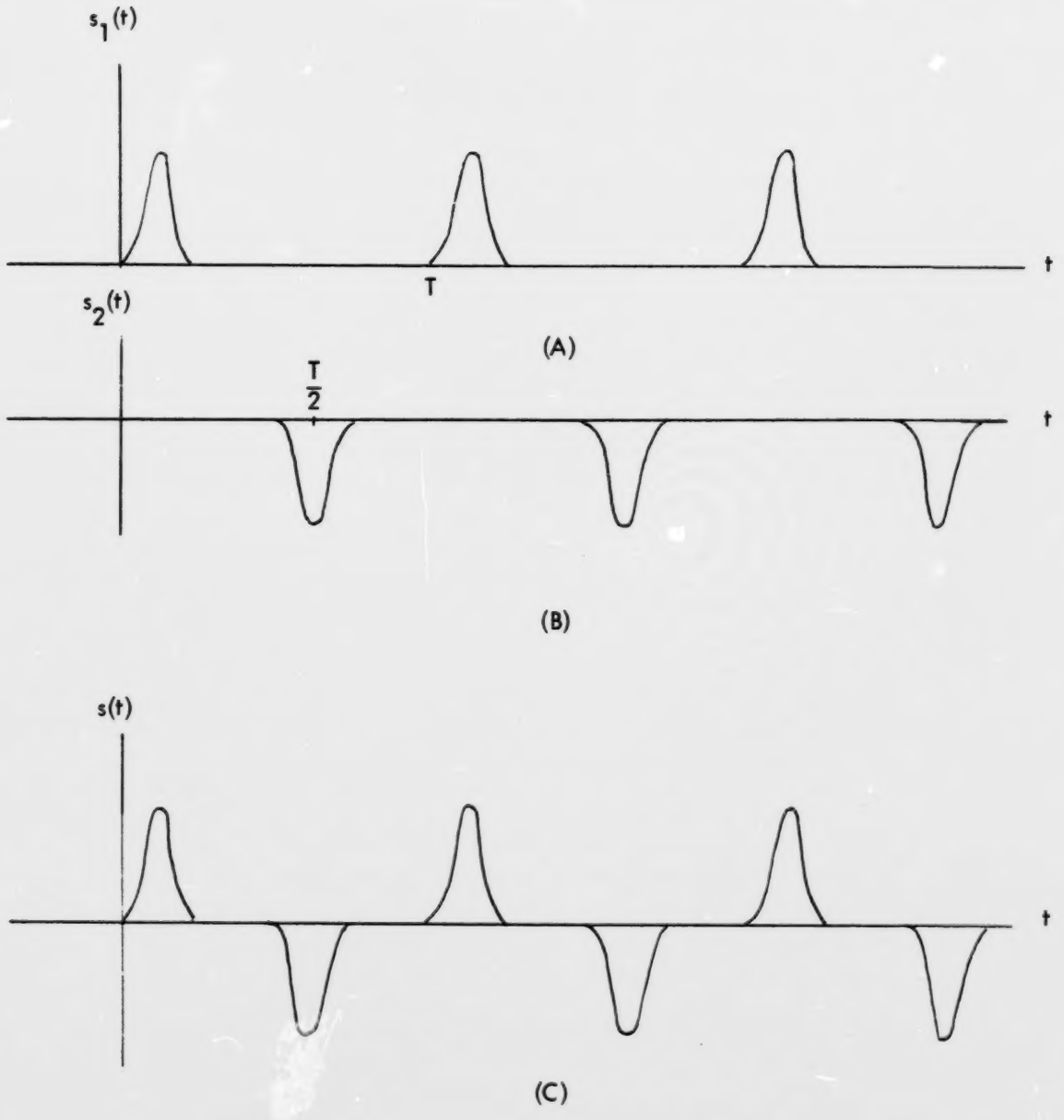


Figure 5.1: Formation of Symmetrical Synthetic Pulse Train with no Even-Numbered Frequencies

Finally, the addition of the two previous trains lead to the bipolar signal of Figure 5.1(C) which can be looked at as a new synthetic pulse train.

This new synthetic pulse train can be shown to contain only the odd-numbered frequencies of the basic pulse train.

From Fourier analysis,

$$s_1(t) = \sum_{1}^{Q} A_{2m-1} \cos\left[(2m-1)\,\omega_o t + \phi_{2m-1} \right] + \sum_{1}^{Q} A_{2m} \cos\left[2m\omega_o t + \phi_{2m} \right]$$

then,

$$s_2(t) = -s_1\left(t - \frac{T}{2}\right)$$

$$s_2(t) = \sum_{1}^{Q} A_{2m-1} \cos\left[(2m-1)\omega_o t + \phi_{2m-1} \right] - \sum_{1}^{Q} A_{2m} \cos\left[2m\omega_o t + \phi_{2m} \right]$$

Now, addition of $s_1(t)$ and $s_2(t)$ leads to

$$s(t) = 2 \sum_{1}^{Q} A_{2m-1} \cos\left[(2m-1)\omega_o t + \phi_{2m-1} \right]$$

clearly, s(t) only contains the odd-numbered frequencies.

This concept is of considerable interest, because it would allow important simplifications in the hardware realization of the RF section. Frequency generation, filtering, amplification and matching would be greatly simplified. Furthermore, signal

digital processing through fast Fourier transform would also be simplified, because it is easier and quicker to handle only the odd frequencies.

This concept will be further investigated during the system analysis phase of the developmental prototype.

5.3 Elimination of the Variable Capacitance Diode Phase Shifters

The control of the relative amplitudes and phases of the synthetic pulse frequency components is necessary to optimize the pulse shape and therefore the sensor resolution. In the experimental prototype, phase shifting has been achieved by using variable capacitance diodes requiring large voltage variations. Furthermore, these devices are complex, because they are made of many varactors to maintain an acceptable level of linearity. For the developmental prototype, it is intended to control the amplitude and the phase of each component frequency through PIN diodes attenuators only. The basic principle of this method is illustrated in Figure 5.3.

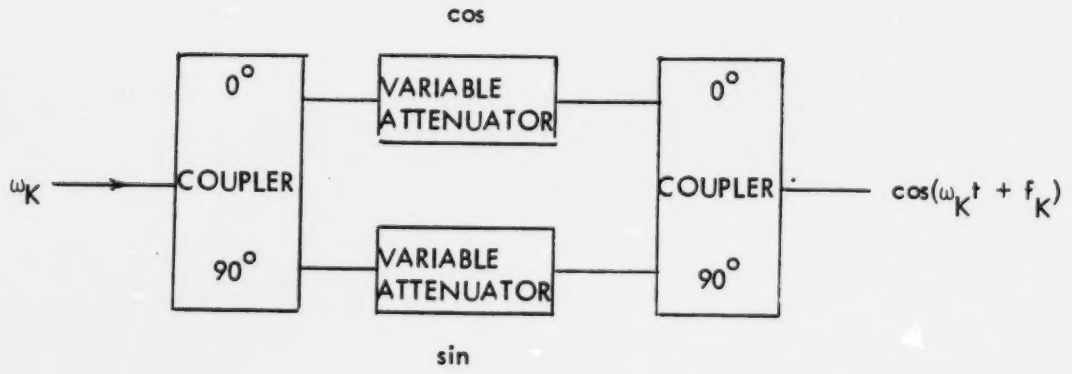


Figure 5.3: Phase Shifting Through Variable Attenuation

The incoming frequency component is split into in-phase (cos) and quadrature (sin) components through the 0° - 90° coupler. These components are then attenuated individually and added in the 0° - 0° coupler to produce the desired amplitude and phase.

6. CONCLUSIONS

The critical elements of the RF section of a fully automated synthetic pulse sea-ice thickness sensor have been identified as the sampling and hold unit, the power amplifier and the receiver RF switch. These elements have been investigated theoretically and experimentally and the following conclusions can be drawn:

- direct periodic sampling of the received signal is preferable to aperiodic or periodic stretched pulse sampling, because it does not impose any restrictions on the sensor operating altitude and therefore does not require any frequency conversions.
- a transmitter peak output power in the 1W to 10W range will be necessary to overcome the background noise encountered when testing the sensor in areas close to VHF transmitters. If a broadband power amplifier is used, its internal harmonic generation level should be lower than - 30 dB, otherwise synthetic pulse adjustment will be made impossible. One way to eliminate this requirement is to use separate narrowband power amplifiers for each frequency component of the synthetic pulse.
- the receiver RF switch of the experimental prototype does not have, in the off state, the power handling capability required by the developmental prototype and the switching transients level is too large. Consequently, a new balanced switch configuration is proposed, using PIN diodes instead of Schottky diodes. The new switch will be equipped with a transient level adjustment scheme and will have increased power handling capability and isolation.

Two ways of reducing the complexity of the RF section have been investigated:

- the first way consists in eliminating the even-numbered frequencies from the synthetic pulse. This method will reduce the complexity of the frequency generation section and decrease the filtering requirements without affecting the sensor resolution.
- the second way aims at eliminating the variable capacitance diode phase shifters used for controlling the relative phase of the synthetic pulse component frequencies. These devices will be replaced by variable attenuators operating on the in-phase and quadrature components of each frequency.

Finally, all the concepts proposed within the frame of this contract will be experimented during the effective development phase of the sensor.

APPENDIX A

MEASUREMENTS OF A SAMPLE AND HOLD UNIT

A.1 SCOPE

The evaluation of a high performance sample and hold unit suitable for aperiodic stretched pulse sampling was carried out primarily to determine the limits of its time and frequency responses. The unit was a ILC Data Devices type DDC-8530.

A.2 TEST LAYOUT

The unit was mounted on a 3" x 4" G40 double-clad PC board. Care was taken to keep logic-level signals as decoupled from the AC and input signals as possible. Ground connections were made so as to minimize coupling of the gate signal into the output signal, while leads and connections (SMA) were arranged to minimize spurious responses and provide the best impedance matching.

A.3 INITIAL ADJUSTMENTS

In order to achieve the greatest accuracy possible, trimming of the DC offset and pedestal were required.

A.3.1 DC Offset

The method used for these initial adjustments consisted of applying a 50 ns negative ECL pulse at a 1 MHz rate to the gate input and adjusting the control for a minimum level reading. The DC offset level fell from 105 mV before trimming to 5 mV after trimming (See Figure A3.1).

A.3.2 Pedestal

The procedure for reducing the pedestal was similar to that used for adjusting the DC offset. A level of 5 mV was measured after adjustment (See Figure A3.1)

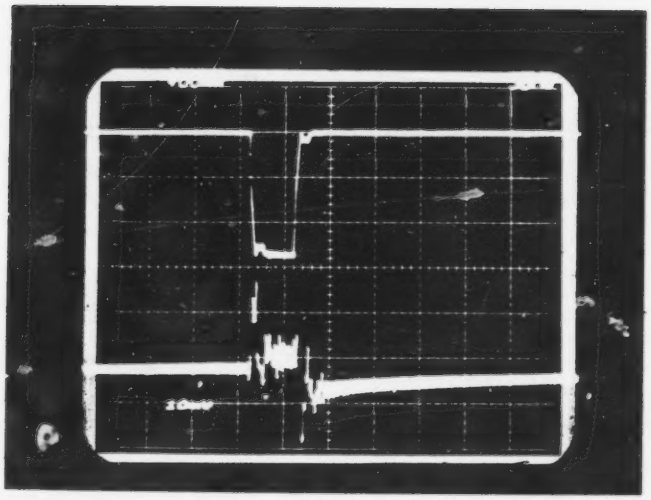


Figure A3.1: Output of Sample and Hold Unit (bottom trace) after Initial Adjustments

Horizontal Scale:	50 ns/div
Vertical Scale:	500 mV/div (top)
	10 mV/div (bottom)

A.4 GAIN AND BANDWIDTH MEASUREMENTS

The objective here was to determine the frequency up to which the unit would still be capable of operating with acceptable attenuation.

A.4.1

The test procedure for these measurements is based on the following set-up:

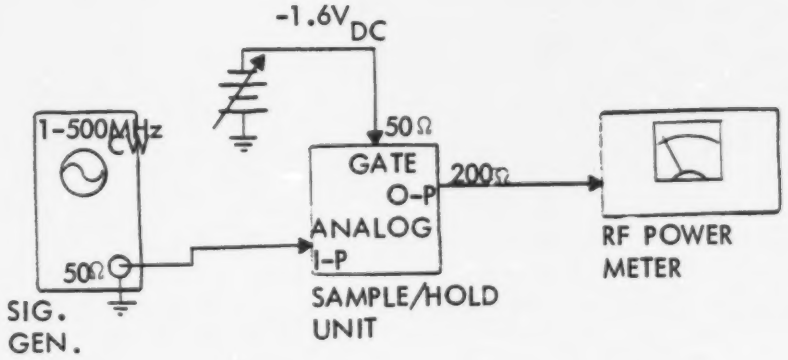


Figure A4.1 Set-up for Gain and Bandwidth Measurements

As the diagram indicates, the unit's internal diode bridge switch was forward-biased ("ON" state) with a constant DC-level (ECL "LO"), while a 20 mW CW signal was applied at the input.

A.4.2 Large Signal Frequency Response

While the large-signal BW was not sepcified for the DDC-8530, it did hold to within limits stated for small-signal BW up to about 40 MHz. The abrupt decrease in gain beyond this point (see Figure A.4.2 and Table A.4.1) shows the frequency limitations of the unit.

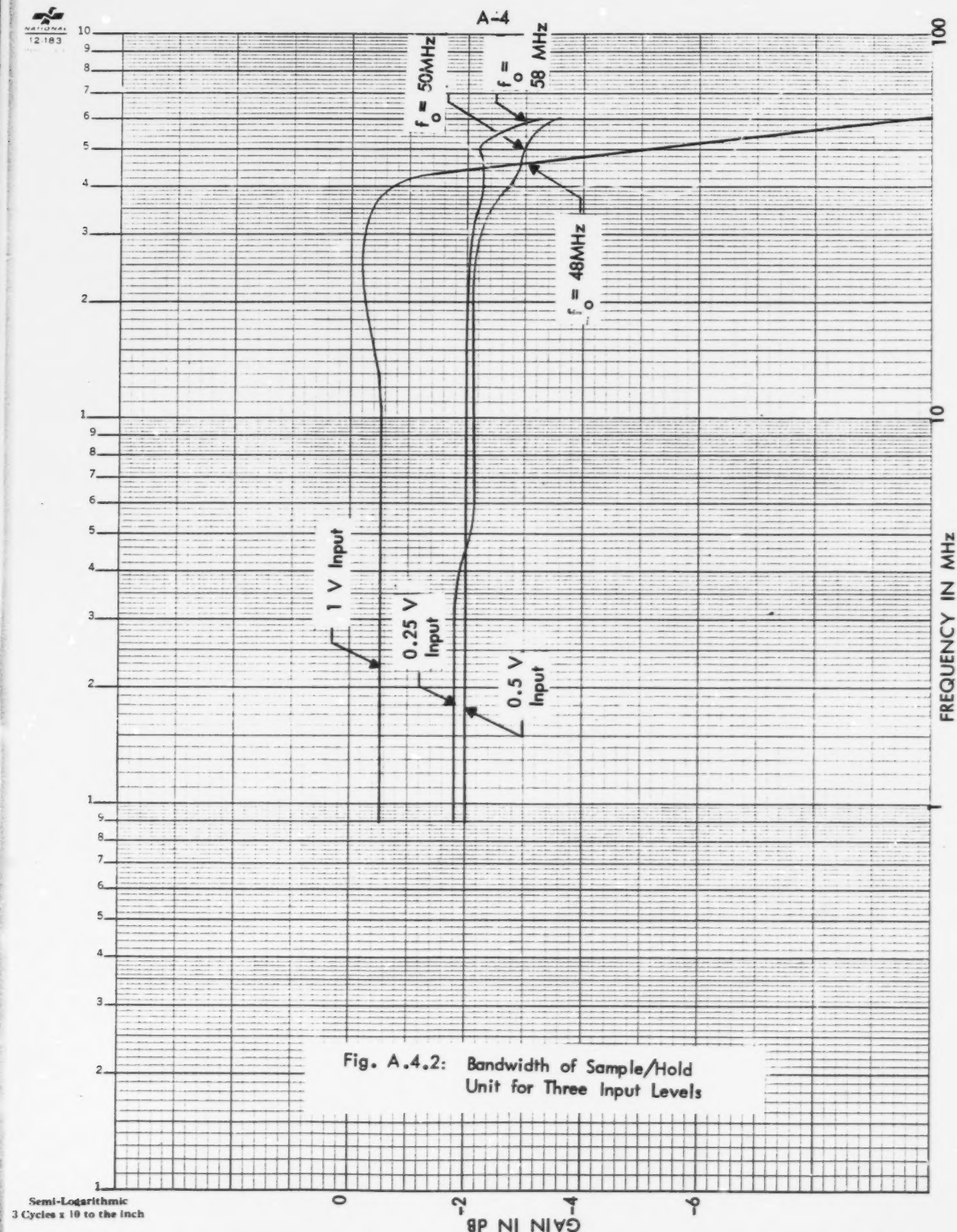


Fig. A.4.2: Bandwidth of Sample/Hold Unit for Three Input Levels

Semi-Logarithmic
3 Cycles x 10 to the inch

Table A.4.1: Large Signal Gain Measurements (Note: Input Signal is 1VRMS or 20 mW)

Input Frequency (MHz)	Gain
1	0.94
10	0.94
20	0.97
40	0.92
50	0.59
60	0.30

A.4.3 Small-Signal Frequency Response

The set-up used was identical to that described in section A.4.2, with the input level changed to 0.25 VRMS (1.25 mW) and 0.5 VRMS (5 mW) successively. Table A.4.2 shows the variation of the units gain with frequency.

Table A.4.2: Small Signal Gain Measurements

Input Frequency (MHz)	Gain dB	
	pi = 1.25 mW	pi = 5 mW
1	-1.8	-2.0
10	-2.1	-2.0
20	-2.1	-2.1
40	-2.7	-2.3
50	-3.0	-2.3
60	-3.6	-3.2

A.5 DYNAMIC TIME RESPONSE EVALUATION

The objectives of these measurements were to evaluate the transient response characteristics of the unit, its sampling capabilities at higher frequencies, as well as the maximum sampling rate at which it can operate.

A.5.1 Rise and Fall Time

Using the test set-up shown below, the rise and fall times of a 1 MHz positive ECL pulse of 50 ns duration were measured at the input and output. (See Figure A.5.2)

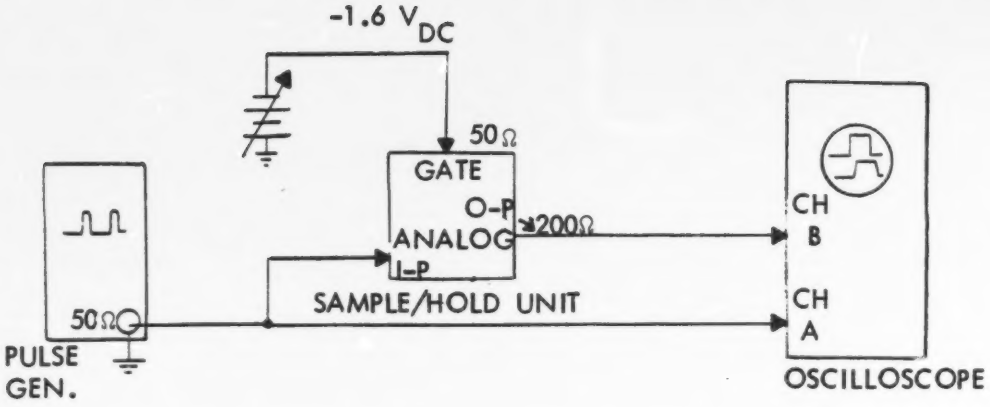


Figure A.5.1: Rise and Fall Time Measurement Set-Up

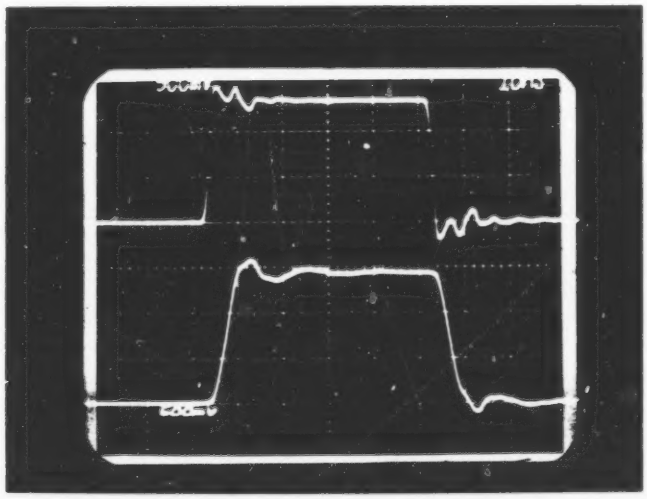


Figure A.5.2: Transient Response of Sample/Hold Unit

(Note: Top Trace - Input, Bottom Trace - Output)

Horizontal Scale: 10 ns/div
Vertical Scale: 500 mV/div (both)

Table A.5.1: Transient Response Measurements

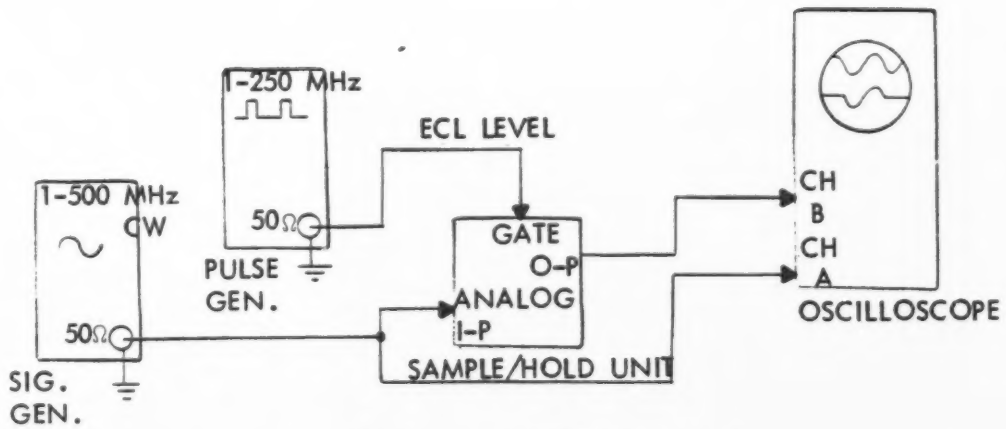
Parameter	Input	Output
t_r	1.5 ns	3.0 ns
t_f	1.0 ns	3.0 ns

A.5.2 Slew Rate

With the same set-up, the slew-rate (rate of rise for high-level pulses) was measured. Its rated 400 V/μs was slightly exceeded (417 V/μs).

A.5.3 Acquisition Time Measurements

The acquisition time is an important parameter which reflects the capability of the unit to sample high frequencies. The test set-up used is shown below together with a sequence of photographs illustrating the sampling action at various input frequencies.



Note: Track = Logic "0" (-1.6 V_{DC})

Hold = Logic "1" (-0.8 V_{DC})

Figure A.5.3: Acquisition Time Test Set-Up

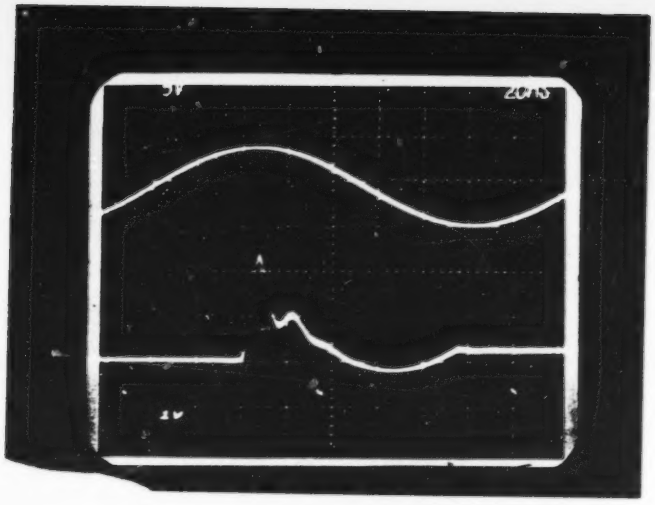


Figure A.5.4: Sampling Action at 5 MHz
$t_{acq} \approx 20$ ns

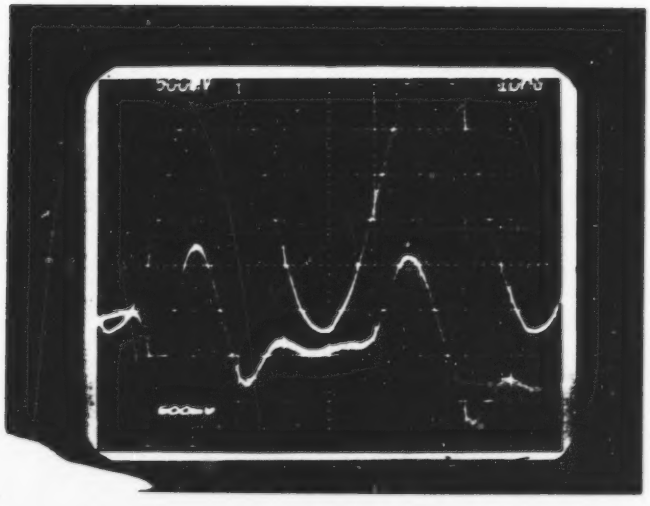


Figure A.5.5: Sampling Action at 20 MHz
$t_{acq} \approx 15$ ns

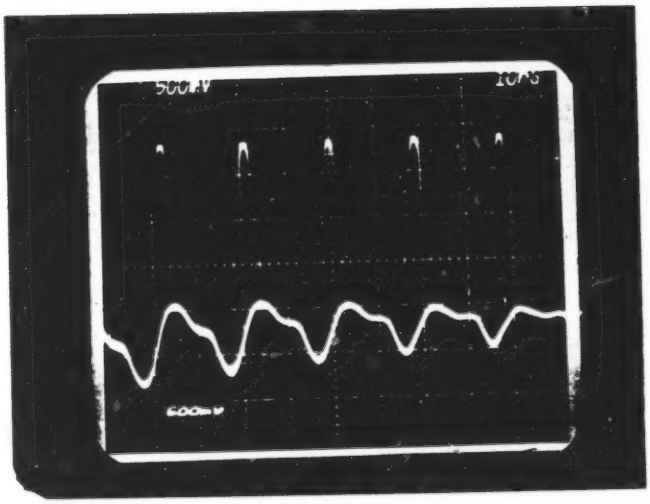


Figure A.5.6: Sampling Action at 50 MHz
$t_{acq} \simeq 12$ ns

A.5.4 Maximum Sampling Rate

The maximum sampling rate is defined by the minimum sample time (acquisition time) and the minimum hold time.

$$\text{Maximum Sampling Rate} = \frac{1}{\text{total cycle time}}$$

$$= \frac{1}{(\text{sampling time})_{min} + (\text{hold time})_{min}}$$

Measured figures of 12 ns (approximate) for both minimum sample and hold times gave a corresponding maximum sampling rate of 41.7 MHz, compatible with the specified 40 MHz limit.

APPENDIX B

MEASUREMENTS OF A POWER AMPLIFIER

B.1 SCOPE

In addition to the evaluation of the amplifier risetime and harmonic generation, the efficiency, bandwidth, and peak power capabilities of the unit were also measured. The sensor experimental prototype was used for most measurements in order to evaluate the performance of the amplifier with complex input waveforms.

B.2 EFFICIENCY (DC to RF) MEASUREMENTS

In the idling state, the amplifier (Microwave Power Devices Inc. type EWAL2015-11) current drain is 4A, corresponding to a DC power consumption of 95W. The DC to RF efficiency is then

$$\frac{P_{RF}}{P_{DC}} = \frac{10W}{95W} = \underline{10.5\%}$$

B.3 FREQUENCY RESPONSE MEASUREMENTS

The objectives of the frequency response measurements were to evaluate the amplifier gain performance for the five frequencies (30 MHz, 60 MHz, 90 MHz, 120 MHz, 150 MHz) used in the sensor experimental prototype.

B.3.1 Test Set-Up

The procedure for measuring the frequency response was based on the following set-up:

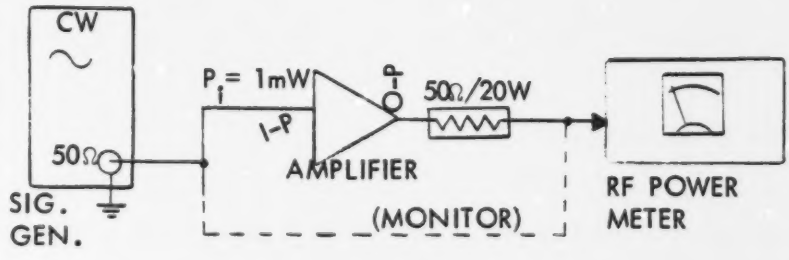


Figure B.3.1: Frequency Response Test Set-Up

B.3.2 Measurements

Table B.3.1 provides the set of measurements recorded from 10 to 300 MHz.

Table B.3.1: Frequency Response Measurements of Amplifier

Input Frequency (MHz)	Po (dBm)
10	+40.00
20	+40.20
30	+40.10
60	+39.75
90	+39.65
120	+39.90
150	+40.35
200	+40.35
300	+39.25

(Note: Input power = 1 mW, or 0 dBm CW
RL = 50 Ω/20W
Po (rated) = 40 dBm = 10W)

Full Power Bandwidth: 10 - 300 MHz ±0.75 dBm

B.4 HARMONIC DISTORTION MEASUREMENTS

To evaluate the harmonic generation of the amplifer, a single tone at 20, 30, 60, 90, 120, 150 and 200 MHz was fed to the input (for rated Po) and the output spectrum observed. Even though the harmonic content of the generator was rather high, results indicated that the amplifier harmonic distortion remained within specified limits, i.e.,

2nd Harmonic: < -22 dB from 20 - 200 MHz (-20 dB rated)

3rd Harmonic: < -12 dB from 20 - 200 MHz (-12 dB rated)

B.5 TIME RESPONSE MEASUREMENTS

For this measurement it was felt that the synthetic pulse itself would be the most useful reference signal.

B.5.1 Test Set-Up

Testing for rise and fall times was made with the following set-up:

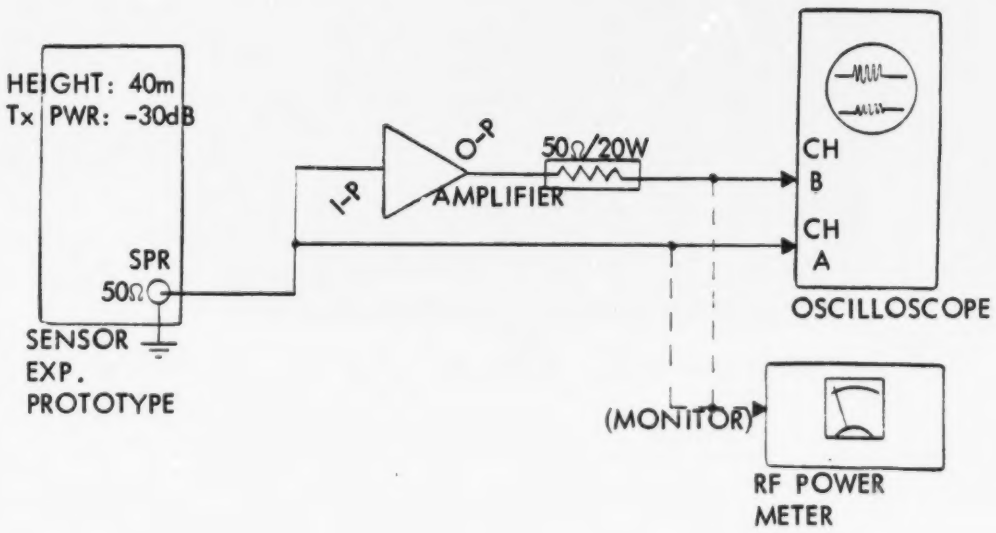


Figure B.5.1: Amplifier Rise and Fall Time Test Set-Up

The synthetic pulse was then optimized to operate at an altitude of 40 m (typical). Figure B.5.2 is a photograph of the synthetic pulse after optimization.

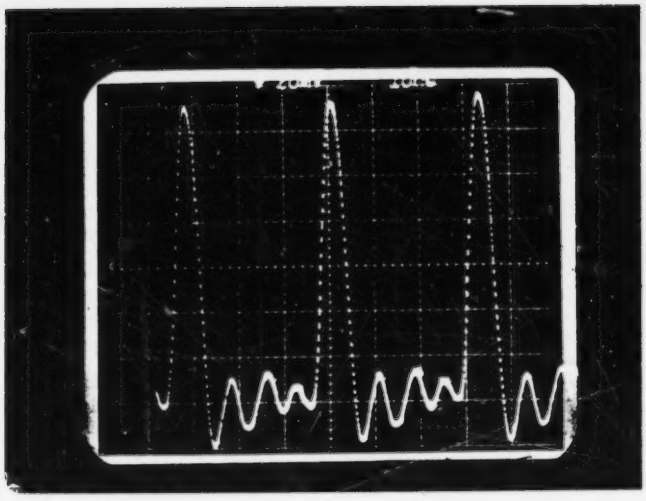


Figure B.5.2: Sensor Experimental Prototype Synthetic Pulse
(Note: pulse duration ≈ 6 ns)

Horizontal Scale: 10 ns/div
Vertical Scale: 20 mV/div

The signal duty cycle was set to be about 40%, i.e. $t_{ON} \approx 440$ ns, $t_{OFF} \approx 600$ ns, and T = 1 μs. The delay from input to output was about 10 ns.

B.5.2 Rise and Fall Times

Both rise and fall time were measured by looking at the envelope and recording the 10 - 90% time elapsed from the "off" level to the positive peak point. Figures B.5.3 and B.5.4 show the portion of the output (amplified) signal used to determine rise and fall times.

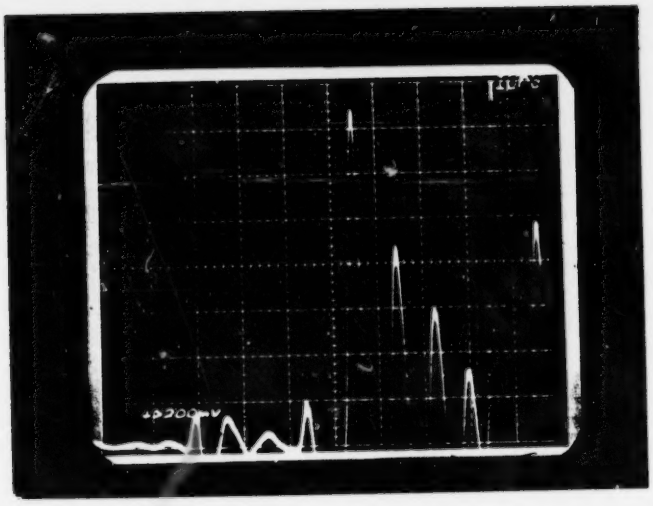


Figure B.5 .3 Output (Amplified) Signal of Amplifier
Note: $t_r \simeq 12$ ns

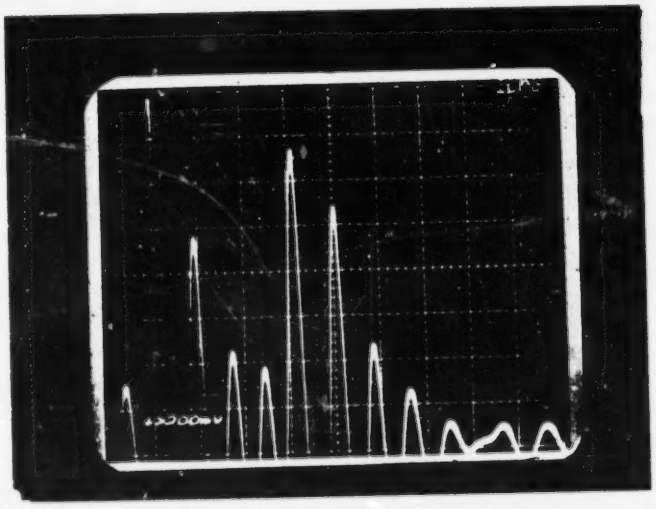


Figure B.5.4: Output (Amplified) Signal of Amplifier
Note: $t_f \simeq 25$ ns

B.6 PEAK POWER MESAUREMENTS

The set-up of Figure B.5.1 was again used for this measurement.

B.6.1 Measurement

Figures obtained through these measurements only provide the average power level, and hence the peak value must be computed from the following relationship:

$$P_p = \left[\frac{\int_0^T p\,dt}{t_{ON}} \right] = \frac{2T\ Pav}{t_{ON}}$$

In the present case, the cycle time was 1.04 μs, and the ON time (signal present) t_{ON} equal to 440 ns. Therefore,

$$P_p = \frac{2 \times 1.04\mu}{440n} \times Pav = 4.73\ Pav$$

Table B6.1 gives the measurements obtained for three input power levels, while Figures B.6.1 to B.6.3 show input-output waveforms for these three cases.

Table B.6.1: Peak Power Measurements of Amplifier

Average Input Power (mW)	Output Power (W)	
	Avg.	Peak
0.15	1.0	4.73
0.85	5.0	23.65
1.60	10.0	47.30

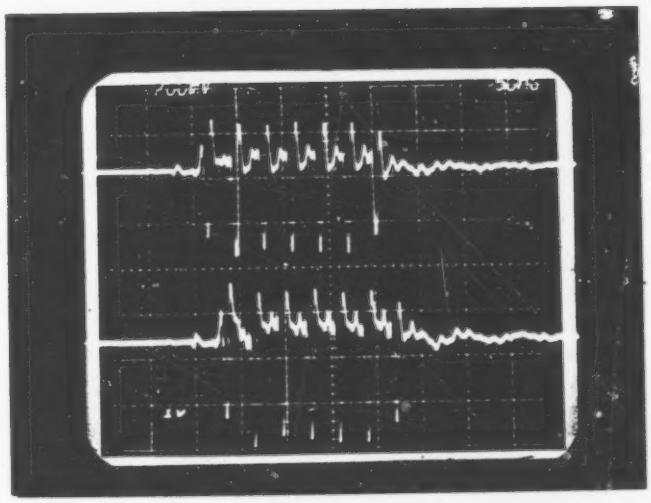


Figure B.6.1: Input/Output Waveforms for Pav = 1W
(P_i (avg) = 0.15 mW)

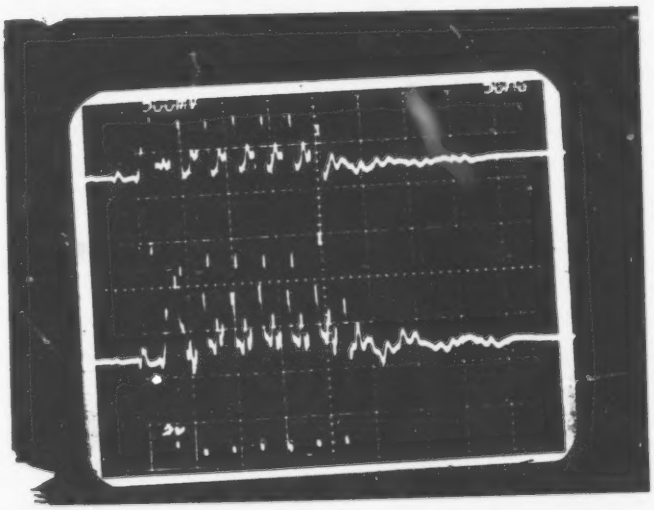


Figure B.6.2: Input/Output Waveforms for Pav = 5W
(P_i (avg) = 0.85 mW)

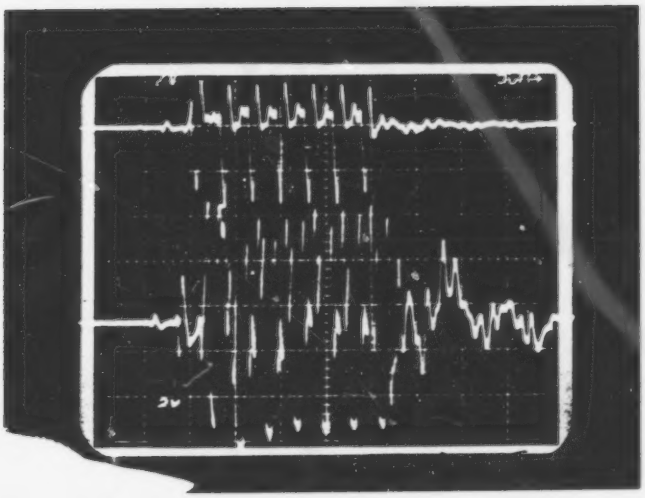


Figure B.6.3: Input/Output Waveforms for Pav = 10W
(P_i (avg) = 1.6 mW)

B.6.2 Input Overload Point

This is the level at which the amplified signal begins to appear severely distorted compared with the input. Overload occurred for an average input power level of about + 10 dBm, which is in agreement with the specified value.

APPENDIX C

MEASUREMENTS OF AN RF SWITCH

C.1 SCOPE

The objective here was to achieve a full characterization of the receiver switch used in the sea-ice thickness sensor experimental prototype and to investigate ways of improving the transient behaviour of the switch in view of its potential use in the developmental prototype.

C.2 TRANSIENT BEHAVIOUR

One of the most critical aspects of the present receiver switch is poor transient performance. A new scheme was implemented to try to improve the behaviour of the basic switch (See section C.2.3)

C.2.1 Test Set-Up

The test set-up used to observe the switch's transient behaviour is shown in Figure C.2.1.

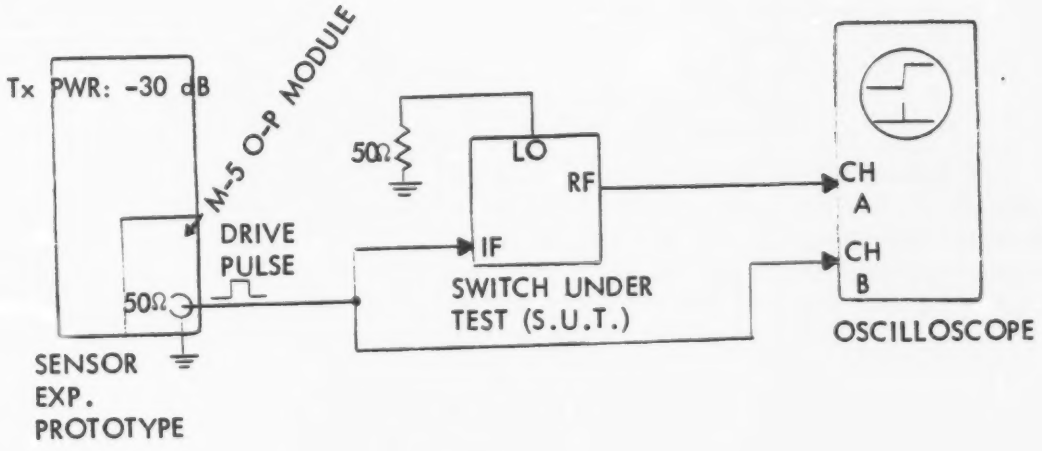


Figure C.2.1 Receiver Switch Transient Behaviour Test Set-Up

C.2.2 Measurements

Figures C.2.2 and C.2.3 show the transients appearing at the RF output at the times of switching "ON" and "OFF".

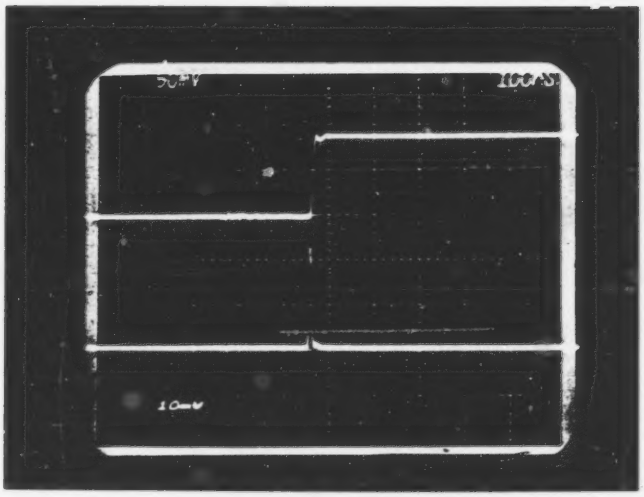


Figure C.2.2: OFF-to-ON State Switching Transient of Receiver Switch
(Note: Top trace: input drive pulse
Bottom trace: RF output in absence of signal
Amplitude of transient: 23 mV)

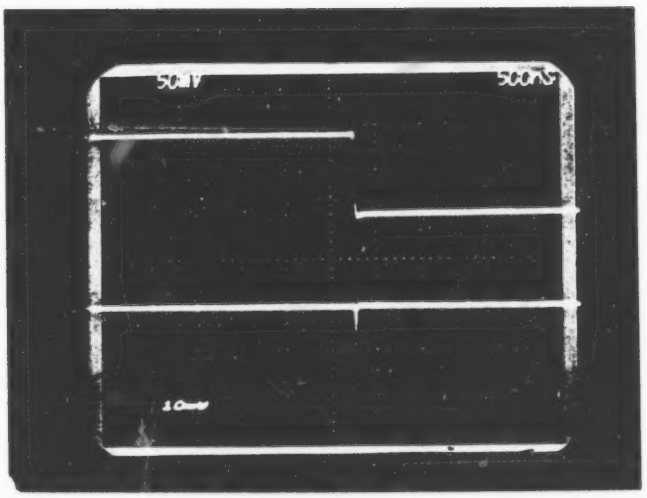


Figure C.2.3: ON-to-OFF state switching
Transient of Receiver Switch
(Note: Amplitude of Transient: 20 mV)

C.2.3 Transient Suppression Scheme

Because of the relatively poor transient performance of the present switch, another configuration was developed whereby adequate attenuation of the transients could be accomplished. This configuration (See Figure C.2.4) uses two identical switches (Minicircuits Laboratory type ZMY-3, as before) driven simultaneously; the transients appearing at the RF output are applied to the in-phase and out-of-phase ports of a hybrid transformer and tend to cancel each other.

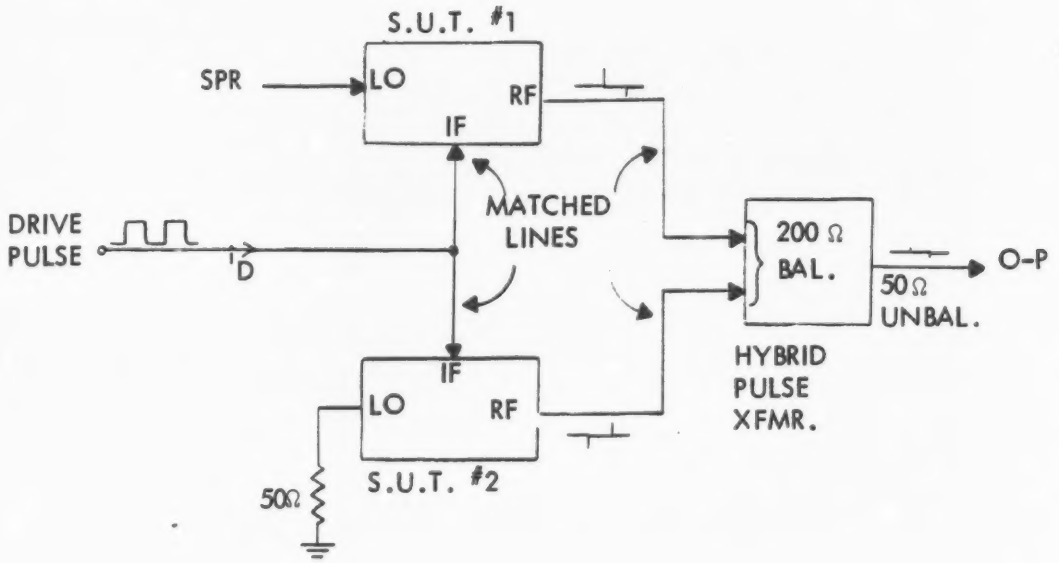


Figure C.2.4: Transient Suppression Scheme

With the sensor experimental prototype set to a height of 40 m and TX power increased by 10 dB, the following photographs were recorded, showing improved transient behaviour.

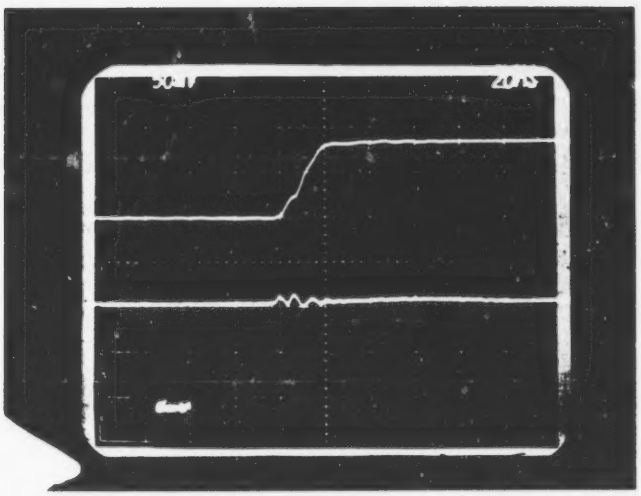


Figure C.2.5: OFF-to-ON Switching
Transient with New Scheme
(Note: Amplitude of Transient: 2 mV)

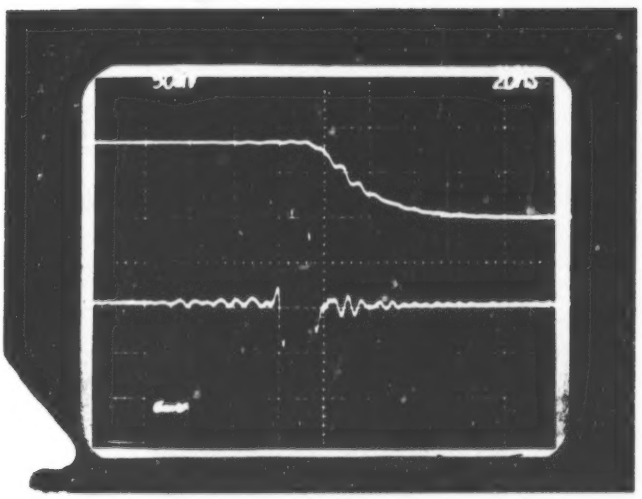


Figure C.2.6: ON-to-OFF Switching
Transient with New Scheme
(Note: Amplitude of Transient: 24 mV)

C.3 INSERTION LOSS MEASUREMENTS

C.3.1 Test Set-Up

The test set-up for measuring the insertion loss of the switch is shown in Figure C.3.1.

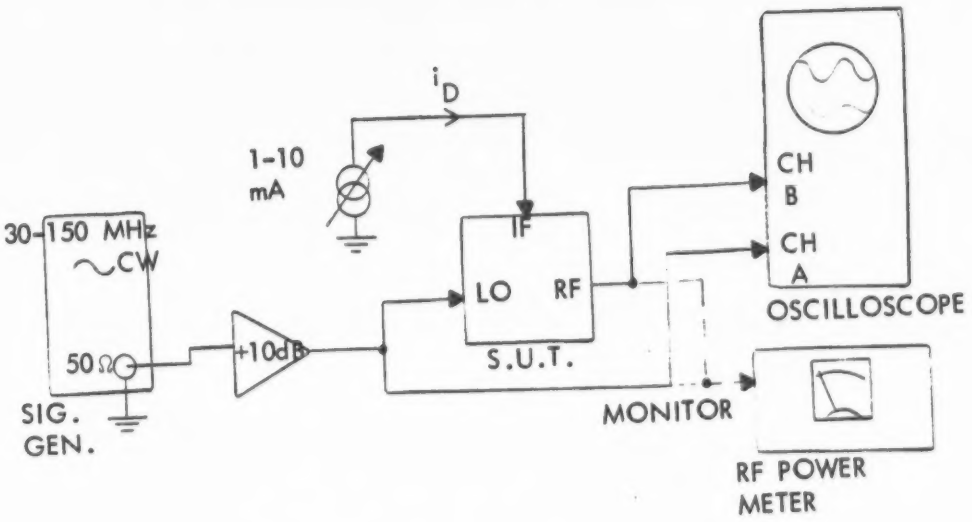


Figure C.3.1: Insertion Loss Test Set-Up

C.3.2 Measurements

Table C.3.1 gives the results before and after substitution of the new configuration.

Table C.3.1: Insertion Loss Measurements

(Note: f_{LO} = 30 to 150 MHz, P_i = 4 mW)

Configuration	i_D (mA)	I.L. (dBm)
original	5	9.26
	10	4.18
new (section C.2.3)	5	10.45
	10	8.11

The switch's turn-on threshold was 0.36V at 60 μA.

C.4 ISOLATION MEASUREMENTS (LO-to-RF)

C.4.1 Test Set-Up

The test set-up for measuring the isolation of the switch is shown in Figure C.4.1.

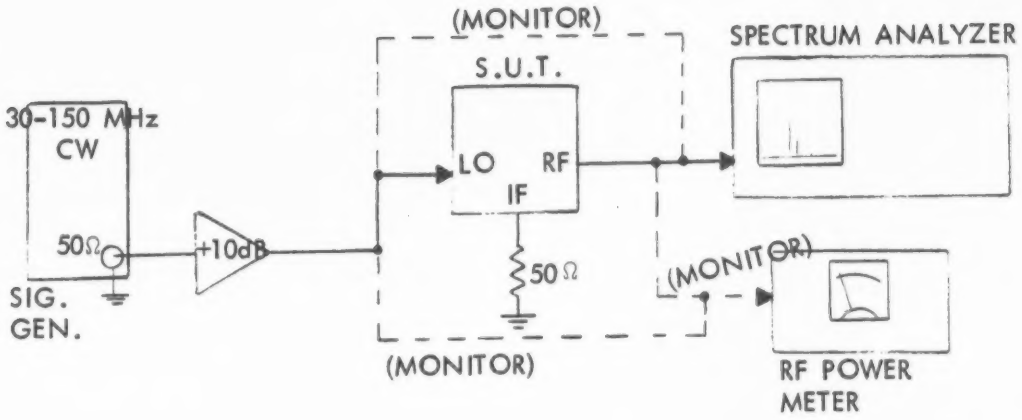


Figure C.4.1: Receiver Switch Isolation Test Set-Up

C.4.2 Measurements

Results are tabulated below for both switch configurations.

Table C.4.1

Configuration	P_i (mW)	Isolation (dB)
original	0.5 4	59.4 62.8
new	0.5 4	61.2 63.2

C.5 INTERMODULATION DISTORTION MEASUREMENTS

C.5.1 Test Set-Up

The test set-up for measuring the intermodulation distortion is shown in Figure C.5.1.

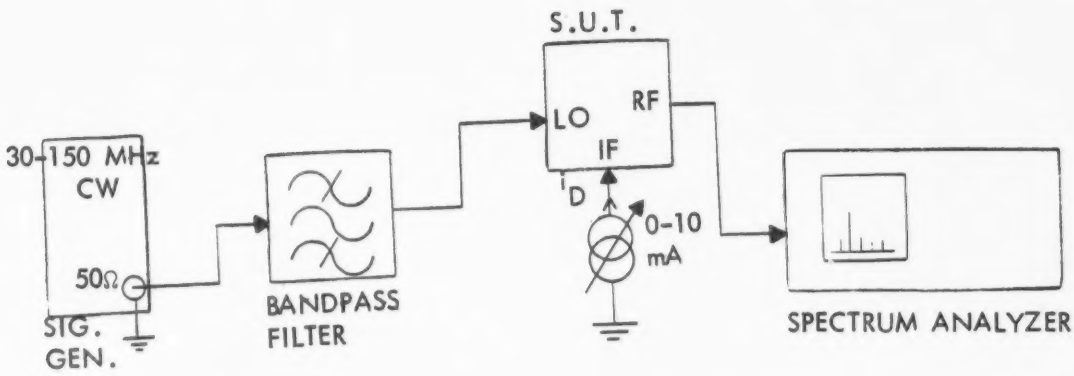


Figure C.5.1: IMD Test Set-Up

C.5.2: Measurements

Controls were adjusted for the following levels:

i_D: 5 and 10 mA

P_i: 0.5, 4.0 and 8.0 mW

f_{LO}: 30, 90 and 150 MHz

Data compiled gave firm evidence of the switch's moderate third-order harmonic contribution. Measurements are shown in Table C.5.1.

Table C.5.1: IMD Measurements of Receiver Switch

f_{LO} (MHz)	P_i (mW)	Third-Order Harmonic (dB)* i_D = 5 mA	i_D = 10 mA
30	0.5	-40	-45
	4.0	-38	-35
	8.0	-32	-36
90	0.5	-40	-48
	4.0	-32	-33
	8.0	-26	-28
150	0.5	-60	-70
	4.0	-40	-64
	8.0	-36	-44

* Relative to 0 dB level at f_{LO}

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